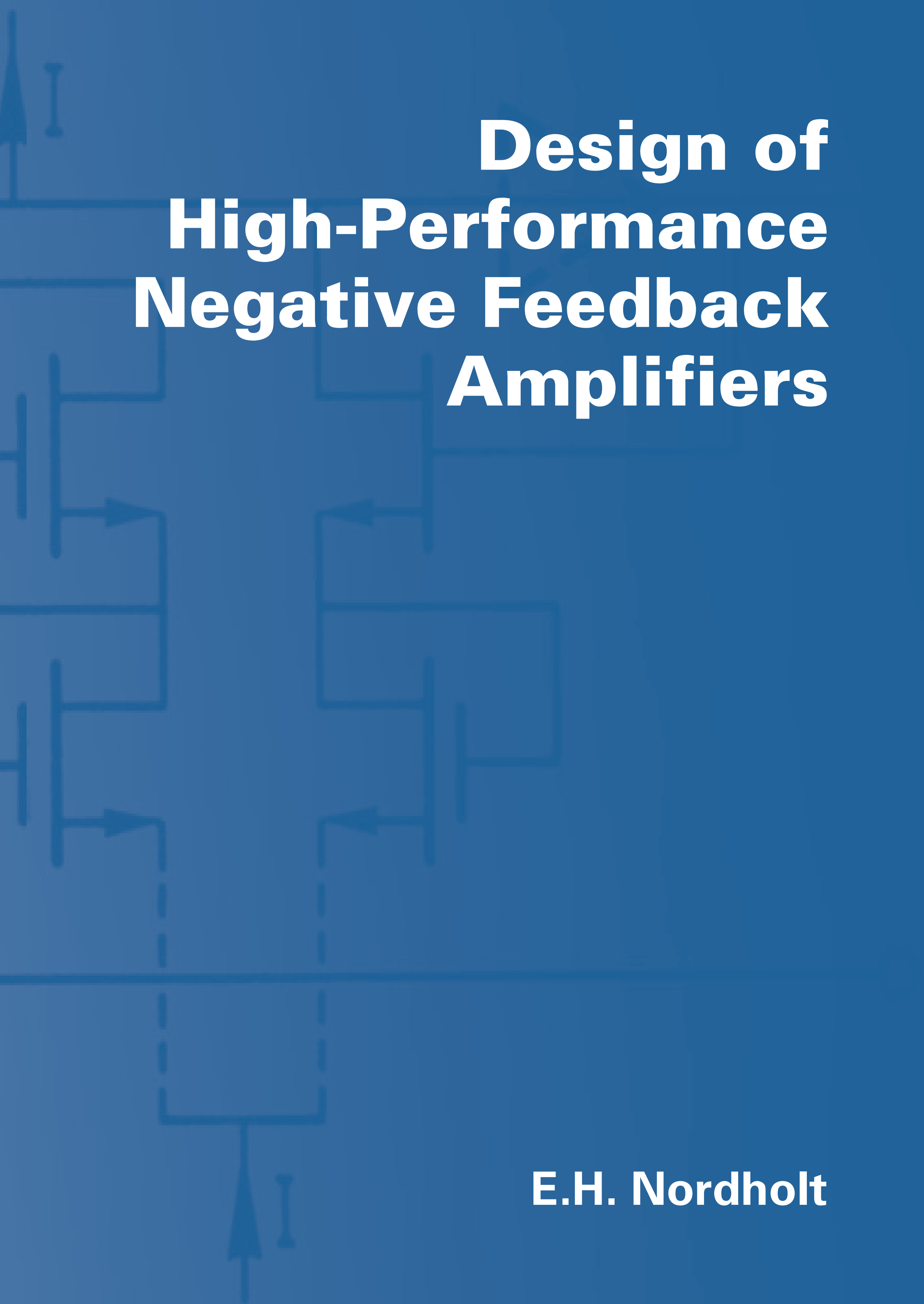

Design of High-Performance Negative Feedback Amplifiers

E.H. Nordholt

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Preface

Amplifier design is very often regarded as making a selection from the large arsenal of known amplifier circuits and then adapting it for a specific purpose, possibly with the aid of computer-aided-design programs. Now and then designers are surprised by the introduction of a new amplifier circuit performing better in some respect than the others.

Each aspiring designer has to find his own way in this jungle. He has to choose from a rather chaotic and scattered collection of amplifier circuits rather than apply a systematic and straightforward design sequence that enables him to design his special-purpose amplifier circuit. A great deal of experience is essential.

This work is an attempt to make a useful contribution to the extensive literature on the subject of amplifier design. It can be justified on the grounds that the approach is believed to be unique in a number of respects. Many works that promise to cover the subject are instead concerned with analysis. Moreover, they frequently deal specifically or separately with particular design aspects, characterised by descriptions such as ‘wide-band’, ‘low-noise’, ‘low-distortion’, etc.

A treatment of the various design aspects and their interconnections, however, is necessary for fruitful amplifier design. At the basis of such a treatment lies the observation — usually easily overlooked — that amplifier design is concerned in the first place with obtaining an adequate quality of information transfer. Amplifiers are more than electronic circuits merely bringing the source power up to a higher level.

Quality requirements are imposed on the signal transfer relative to the type of information and to the manner of perception, registration, or processing. The quality of information transfer is determined by a large number of quality aspects such as linearity, accuracy, efficiency, signal-to-noise ratio, etc. Unfortunately, it cannot be expressed as a quantitative figure of merit

Trade-offs between various quality aspects are likely to emerge. Sometimes they will be of a fundamental nature and are imposed by physical and technological limitations, but frequently they will result from the nature of a specific amplifier circuit itself, which — on second thoughts — might not be the most appropriate one to fulfil the desired function.

A systematic, straightforward design approach is presented in this work. It is more or less inspired by the work of Cherry and Hooper[1] which I consider one of the finest design treatises. The present work is more concerned with basic design considerations. Preference was given to a qualitative rather than to a quantitative approach.

Finding the proper configurations for the basic amplifier and of the amplifier stages is considered of primary importance and is emphasised here. This book is therefore largely concerned with the design phase preceding the phase in which existing computer aids can be helpful.

The approach is characterised best by describing it as a systematic and consistent arrangement of design considerations regarding various quality aspects of information transfer. Via the classifications of amplifier configurations, a systematic design method for negative-feedback amplifiers is developed.

A short description of the main lines along which the design method has been developed is given below.

In chapter 1, criteria are given for the adaptation of the input and output impedances to the source and the load, respectively (usually transducers). The purpose of these adaptations is the realisation of optimum information transfer from the signal source to the amplifier and from the amplifier to the load. Next, criteria are deduced for optimum information transfer of the amplifier, preserving signal-to-noise ratio and efficiency by the application of feedback. Classifications are given of basic amplifier configurations with up to four negative-feedback loops, providing the designer with the complete set of fundamentally different two-port amplifier types. The characteristic properties and the practical merits of these configurations are discussed.

A similar classification is given in chapter 2 for configurations with a single active device. A uniform description of these single-device configurations will appear to be of great help in finding suitable stage configurations in the active part of an amplifier with overall negative feedback. The balanced versions of these single-device stages are mentioned but not studied in detail.

In chapter 3, design criteria regarding random noise are formulated. These criteria relate mainly to the selection of the most favourable input-stage configuration and the active device to be used in this stage for a given signal source.

In chapter 4, those configurations of amplifier stages that are best suited to realising optimum accuracy and linearity of information transfer are arrived at.

Thereafter, bandwidth and stability considerations are taken into account in chapter

---

[1] E.M. Cherry and D.E. Hooper, Amplifying Devices and Low-Pass Amplifier Design, John Wiley and Sons, New York, 1968.

5. The requirements for optimum performance in this respect fortunately appear to be to a large extent compatible with the requirements regarding the other quality aspects.

The design of bias circuitry is considered in chapter 6. It will be shown that this part of the design can be done in such a way that the signal-path performance of the amplifier is scarcely affected.

Finally, an outline of the design method is given in chapter 7. For examples of amplifiers designed according to the design procedure developed in this book reference is made to the literature.

This book is a revised and reviewed version of my Ph.D. thesis, which was published in June 1980 under the supervision of Prof.Dr.Ir. J. Davidse. The results of the work that he has encouraged me to carry out in the Laboratory of Electronics at the Delft University of Technology, The Netherlands, can be found here.

Writing such a thesis is perhaps even more of a burden to those in an author's environment than to himself. Without the aid of many others, it would not be possible to obtain a Ph.D. degree. As an acknowledgement of their support, this book is dedicated to all who contributed in some way.

Susan Masotty reviewed the text and corrected my numerous linguistic errors. Wim van Nimwegen drew the figures. Josette Verwaal and Hilda Verwest typed the manuscript. They thus contributed to the mere physical existence of this book, which happens to bear my name. This may, however, veil the fact that an author is no more than a person who is lucky enough to be able to write down the ideas he would never have had without a stimulating environment.

It is to this environment — which usually cannot be adequately indicated by names but includes the above — in all its abstraction that I am deeply indebted.

Delft, 1983 Ernst Nordholt

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1

Basic Amplifier Configurations for Optimum Transfer of Information from Signal Sources to Loads

1.1 Introduction

One of the aspects of amplifier design most treated — in spite of its importance — like a stepchild is the adaptation of the amplifier input and output impedances to the signal source and the load. The obvious reason for neglect in this respect is that it is generally not sufficiently realised that amplifier design is concerned with the transfer of signal information from the signal source to the load, rather than with the amplification of voltage, current, or power.

The electrical quantities have, as a matter of fact, no other function than representing the signal information. Which of the electrical quantities can best serve as the information representative depends on the properties of the signal source and load. It will be pointed out in this chapter that the characters of the input and output impedances of an amplifier have to be selected on the grounds of the types of information representing quantities at input and output.

Once these selections are made, amplifier design can be continued by considering the transfer of electrical quantities. By speaking then, for example, of a voltage amplifier, it is meant that voltage is the information representing quantity at input and output. The relevant information transfer function is then indicated as a voltage gain.

After the discussion of this impedance-adaptation problem, we will formulate some criteria for optimum realisation of amplifiers, referring to noise performance, accuracy, linearity and efficiency. These criteria will serve as a guide in looking for the basic amplifier configurations that can provide the required transfer properties. The suitability of some feedback models will be discussed. The rather unusual, so-called asymptotic-gain model, will be selected for use in all further considerations.

Thereafter, we will present a classification of basic negative-feedback configurations. First, a rather theoretical approach is given, where the active amplifier part is considered as a nullor, while the feedback network is realised with ideal transformers and gyrators. A class of *non-energic* negative-feedback amplifiers results, with up to maximally four overall feedback loops, forming the complete set of fundamentally different two-port amplifier types. More practical configurations with transformer feedback are treated next. Configurations with passive components (except transformers) in the feedback network are investigated in addition. These impedance feedback configurations are the most familiar amplifier types unfortunately not capable of realising all types of desired transfer functions. Therefore, amplifiers with active feedback are considered separately. These can provide the missing types of transfer functions. Two basically different types of active feedback can be distinguished. One of them will referred to here as indirect feedback.

These indirect-feedback amplifiers are especially suitable for realisation in integrated-circuit technology. As a practical restraint it is assumed throughout the whole chapter that source, load and amplifier have one common terminal ('ground').

1.2 General considerations

In figure 1.1 a general representation is given of a signal chain, where an amplifier is inserted between a signal source and a load. The input and output quantities of the source and load, respectively, are not necessarily electrical quantities. The three blocks in figure 1.1 can each be represented by either active or passive twoports. The parameters describing the transfer properties of these twoports will generally be functions of frequency, signal amplitude, temperature and other environmental circumstances. In order to realise an optimum information transfer from source to load, the transfer of each twoport should be determined by the parameter(s) that yield(s) *the best reproducing relation* between an input and an output quantity.

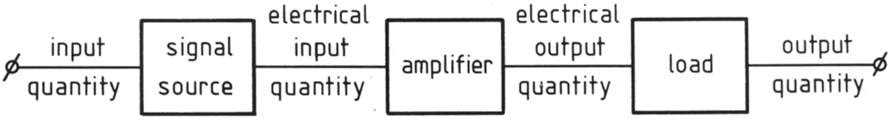


Figure 1.1 Amplifier inserted between signal source and load; source and load are formed by other electronic circuits or input and output transducers.

The best reproducing relation is determined in the case of transducers by their construction, their physical operation mechanism and — in the case of active transducers — by the way auxiliary power is supplied. The best reproducing input-output relation dictates the required character of the load impedance in the case of an input transducer (signal source) and of the source impedance for an output transducer (load).

An illustrative though simplified example of an input transducer is given below. A piezo-electric transducer produces a charge q_s linearly proportional to the pressure p, which is assumed to be the information representing primary signal. The best reproducing relation T of input and output quantities is therefore given by:

$$T = \frac{q_s}{p}$$

The output impedance of this transducer (the source impedance from the viewpoint of the amplifier) can be represented in the low-frequency region by a capacitance C_s. The open output voltage is given by:

$$u_{so} = \frac{q_s}{C_s}$$

This relation between voltage and charge is inaccurate, non-linear and temperature dependent, because of the properties of C_s. The open voltage does not therefore reproduce the primary information in an optimum way. The relation between short-circuit current and charge, however, is given by a linear differential equation:

$$i_{ss} = \frac{\mathrm{d}q}{\mathrm{d}t}$$

Therefore, a very low input impedance of the amplifier is required for optimum information transfer from transducer to amplifier. In other words, no voltage is allowed to arise across C_s for the transfer not to be affected by the less favourable properties of this source impedance.

Input transducers are preferably represented by Thévenin or Norton equivalent circuits. In that case, we have to deal with voltage and current sources only. The equivalent circuit should preferably be selected in accordance with the best reproducing relation of the transducer. In the case of the piezo-electric transducer, the Norton equivalent circuit is the obvious representation. It is shown in figure 1.2, where the current source is related to the charge.

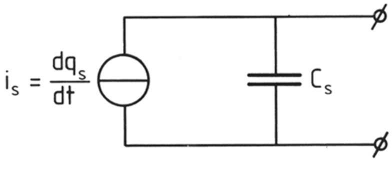


Figure 1.2 Preferable equivalent circuit for a piezo-electric transducer (in the low-frequency region).

If an accurate and linear relationship exists between open voltage and short-circuit current, as for example in *characteristic-impedance* systems, there is no preference for either one of the equivalent circuits. In that case *power* may be the information-representing quantity, and a possible criterion for optimum information transfer is that no power be reflected. To meet this criterion, an accurate and linear input impedance is needed.

Similar considerations can be given for loads. For instance, if the output quantity of the output transducer (load) has a linear and accurate relationship with the driving voltage, the amplifier should behave like a voltage source.

The transfer function of the amplifier which reproduces the information generated in the source as information supplied to the load will be labelled the *transmittance* of the amplifier. The various types of required transmittances are classified in table 1.1. Current, voltage and power are regarded as information-representing quantities. They can be derived quantities, as will be clear from the example of figure 1.2. To meet the requirements imposed by source and load, as far as the information transfer to and from the amplifier is concerned, it is necessary to give the input and output impedances their proper values, which may be either very large, very small, or accurate and linear. Various methods are available for realising the desired input and output impedances. They will be discussed in subsequent sections.

Table 1.1 Classification of different types of transmittances.

information-representing quantity produced by the source	information-representing quantity to be supplied to the load	transmittance
1. voltage	voltage	u_ℓ/u_s (voltage gain)
2. voltage	current	I_ℓ/U_s (transadmittance)
3. current	voltage	U_ℓ/I_s (transimpedance)
4. current	current	I_ℓ/I_s (current gain)
5. voltage	power	P_ℓ/U_s
6. power	voltage	u_ℓ/P_s
7. power	current	I_ℓ/P_s
8. current	power	P_ℓ/I_s
9. power	power	P_ℓ/P_s (power gain)

1.2.1 Preliminary criteria for the realisation of high-quality transmittances

Some general criteria on which the design of high-quality amplifiers has to be based are implicitly formulated below. Supporting material will be given in later chapters.

1. The insertion of series and shunt impedances in the signal path of an amplifier — causing a loss of available signal power — must be avoided. At the input, such a loss results in a relatively enlarged noise contribution of the active devices, while the resistors used add thermal noise to the signal. At the output, it results in a loss of efficiency and possibly increased non-linearity, because the active devices have to handle larger signal amplitudes.

On these grounds, methods for realising sufficiently high or low input or output impedances with the aid of passive series of shunt impedances will be called 'brute force' methods. A consistent application of the above formulation is one of the most important elements of this work.

2. To meet the requirements for optimum information transfer, the transmittance must be made independent of the properties of the active devices, and therefore the application of negative feedback is imperative. Power gain, produced by the active devices, has to be used as the expedient to realise both the desired *magnitude* of the transmittance and the desired *quality* of information transfer.

In this chapter we will deal with the basic negative-feedback amplifier configurations. These configurations can comprise one or more active devices in their *active parts*. When one active device is involved, we will call the configuration a *local-feedback* stage. When more stages are incorporated in the active part, it will be said that the configuration has *overall feedback*.

The loop gain in a multistage overall-feedback amplifier can be much larger than the loop gain in local-feedback stages. As a matter of fact, source and load impedances frequently have such characteristics that it will be impossible to realize the desired transmittance with a single active device. A cascade of local-feedback stages may then be considered, or alternatively, an overall feedback amplifier.

In this and the following chapters it will always be assumed that the feedback loop embraces as many stages as is consistent with optimum performance with respect to all quality aspects to be dealt with. High-frequency performance sets a limit to the useful magnitude of the loop gain and the number of stages in the active part. For solving certain amplifier problems it may be necessary to use two or more cascaded overall-feedback amplifiers. In that case, the designer has the freedom to choose the characters of the first-amplifier output impedance and the second-amplifier input impedance. Cherry and Hooper [1] have shown that, in order to avoid undesired interaction, an impedance mismatch between the amplifiers is favourable. We will not deal with this aspect in this work any further.

For the design of negative-feedback amplifiers, it is desirable to have at one's disposal a feedback model which closely conforms to the above view. In the next section we will briefly discuss some methods for the analysis of feedback amplifiers and the suitability of these methods for design purposes.

1.3 Models for negative-feedback amplifiers

1.3.1 Black's feedback model

Though not necessary, it is frequently considered appropriate to make the presence of negative feedback in an amplifier explicit. For design purposes, one can think about negative feedback as a means of exchanging available power gain for quality of information transfer. This view comes somewhat to the fore in the *elementary feedback model* of Black's feedback patent [2] given in figure 1.3. This model is commonly used as a design aid. The transfer function is given by:

$$A_t = \frac{E_\ell}{E_s} = \frac{A}{1 - Ak} \qquad (1.1)$$

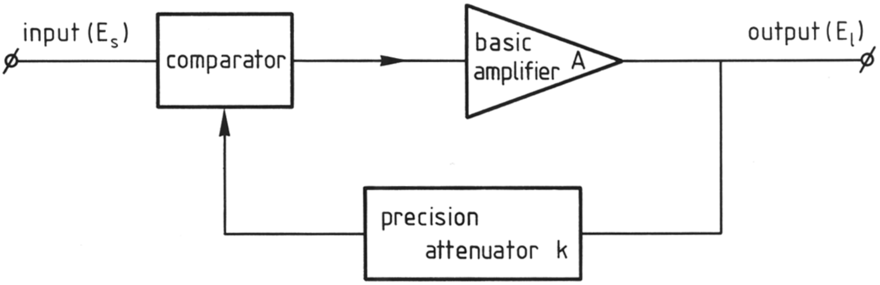


Figure 1.3 Elementary feedback model.

Because the model is a block diagram based on transfer functions which are assumed to be unilateral and not affecting each other, it is hardly applicable to practical amplifier configurations. The calculation of the transmittance of a real amplifier is therefore usually performed with the aid of matrix-parameter representations of the amplifier and feedback network [3], [4]. As a result, an identification problem arises when the expression for A_t found in this way is manipulated into the form of relation (1.1).

As a matter of fact, this method is suitable for analysis rather than for straightforward design. It is applicable to a given configuration but gives no explicit indications for the selection of a configuration for a certain purpose. Besides, the use of this model bears the risk that one might try and realise the transfer functions A and k in such a way that they are indeed unilateral and not affecting each other. As a design philosophy for general purpose operational amplifiers, this may be a defensible approach, but as will become apparent from the subsequent chapters, it does not lead to optimum special-purpose amplifier configurations.

1.3.2 Anti causal analysis

Recently, Waldhauer [5] has proposed the so-called *anti causal analysis* of feedback amplifiers. This analysis is made with the aid of the *transmission* matrix, where the input quantities are expressed as functions of the output quantities. The reciprocal value of the amplifier transfer function is found by this method to be the sum of a number of terms. The most important term describes the design goal and is determined by the components of the feedback network. The other terms express the non-ideality of the active part. Though the presence of feedback is not made explicit, the concept of feedback is used to find the basic configuration for a certain application. This method is, like the previous one, suitable for analysis rather than for design.

1.3.3 Superposition model

An equally formal approach of feedback-amplifier analysis leads to a model that can be indicated as the *superposition model*, because it is derived by using the superposition principle. This model will be derived here, because it forms the basis for the model that will appear to be most suitable for our design purposes.

Each voltage and current in the amplifier can be written as a linear combination of the quantity produced by the signal source and the current or voltage of an arbitrarily selected controlled source in the amplifier. For setting up the network equations, the controlled source is initially regarded as independent. If we indicate the quantities regardless of their dimensions by the symbol E, the network equations for the load and input quantities can be written as:

$$E_\ell = \rho E_s + \nu E_c \tag{1.2}$$

$$E_i = \zeta E_s + \beta E_c \tag{1.3}$$

The indices ℓ, i, s and c refer to load, input, signal source and controlled source, respectively. Besides, there is a relation between E_c and E_i. that can be written as:

$$E_c = AE_i \tag{1.4}$$

A non-zero value of the transfer function β indicates the existence of feedback. The transfer functions ρ, ν, ζ, A and β can have dimensions [Ω], [1/Ω] or are dimensionless. The transfer function A is called the *reference variable*.
The transfer function A_t of the amplifier is found from (1.2), (1.3) and (1.4):

$$A_t = \frac{E_\ell}{E_s} = \rho + \nu\zeta\frac{A}{1 - A\beta} \tag{1.5}$$

The product $A\beta$ is called the *loop gain* with respect to the reference variable A, while $1 - A\beta$ is the well-known *return difference,* with respect to the reference variable A, as defined by Bode [6].
An example may elucidate how A_t is calculated. Figure 1.4 shows an arbitrarily chosen, very simple configuration, where the amplifier is represented by a current source controlled by the input voltage. The transconductance g is selected for obvious reasons as the reference variable. Calculating now the factors

$$\rho = (U_\ell/I_s)_{I_c=0}, \quad \nu = (U_\ell/I_c)_{I_s=0}, \quad \zeta = (U_i/I_s)_{I_c=0} \quad \text{and} \quad \beta = (U_i/I_c)_{I_s=0}$$

we find for the transfer function A_t:

$$A_t = \frac{U_\ell}{I_s} = \frac{Z_i Z_\ell - g Z_f Z_i Z_\ell}{Z_\ell + Z_f + Z_i + g Z_i Z_\ell}$$

If the loop gain $A\beta = -gZ_iZ_\ell/(Z_\ell + Z_f + Z_i)$ is very large, A_t can be approximated by:

$$A_t = -Z_f$$

The transfer function stabilised by the negative feedback in this case is a transimpedance.

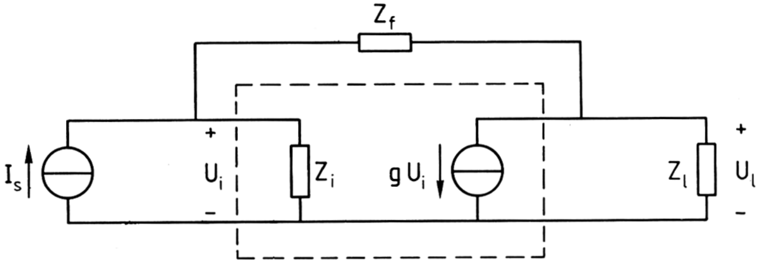


Figure 1.4 Example of a feedback configuration.

Although this superposition model offers the possibility to exactly calculate the transmittance in a relatively simple way, in contrast with other models or calculation methods, once again it is only suitable for analysis. Some qualitative design indications can be derived from the transfer functions ρ, ζ and ν. The direct transfer from source to load is described by ρ, and ζ and ν are measures for the signal loss at input and output, respectively. The model does not, however, provide any explicit indications for the realisation of high-quality transmittances. We have paid so much attention to it, because the so-called *asymptotic-gain model* can easily be derived from it and because the latter model does not have the disadvantages mentioned earlier.

1.3.4 Asymptotic-gain model

(i) The transfer functions

If in (5) the loop gain becomes infinite ($A\beta \to \infty$), the transfer function is given by:

$$\lim_{A\beta \to \infty} A_t = A_{t\infty} = \rho - \frac{\nu\zeta}{\beta} \tag{1.6}$$

where A_t will be referred to as the *asymptotic gain*. Substitution of (1.6) in (1.5) yields:

$$A_t = \frac{E_\ell}{E_s} = \frac{\rho}{1 - A\beta} - A_{t\infty}\frac{A\beta}{1 - A\beta} \tag{1.7}$$

In all practical amplifier designs, the first term in (1.7) will be much smaller than the second, and (1.7) can be written as:

$$A_t = \frac{E_\ell}{E_s} = A_{t\infty}\frac{A\beta}{1 - A\beta} \tag{1.7}$$

The design of the transfer function of a negative-feedback amplifier is thus reduced to two successive steps. The first step is the determination of $A_{t\infty}$, which can be considered as the design goal, and the second step is the realisation of an adequate loop transfer function $A\beta$.

Unless the input and/or output impedances need to be accurate, it is not necessary to calculate their values. The amplifier performance is adequately described by expression (1.8). Note that the transmittance is defined as the transfer function:

$$A_t = \frac{\text{load quantity}}{\text{source quantity}} \qquad \text{and not} \qquad \frac{\text{output quantity}}{\text{input quantity}}$$

To ascertain the meaning of $A_{t\infty}$ we consider the expression for E_i, which follows from (1.3) and (1.4):

$$E_i = \frac{\zeta}{1 - A\beta} E_s \tag{1.9}$$

Obviously, E_i approaches zero when $|A\beta/\zeta|$ becomes infinite. If for E_i the current or voltage at the input of the first amplifier stage is selected, E_i will be constrained to be zero when the loop gain becomes infinite. If, for example, E_i is the voltage between two nodes connected by one branch (as in the example of figure 1.3) then the current in that branch will also approach zero when $A\beta$ becomes infinite. The constraints correspond to the *nullor* [7] constraints in network theory.

(ii) Impedance calculation

The superposition principle can also be used to calculate input and output impedances of the amplifier. Only in the case where these impedances are intended to become accurate and linear by the feedback action will the calculation be useful. We will follow here the treatment as given by Boon [8].

Instead of E_ℓ, which is the load quantity in the case of the gain calculation, we will in the network equations now write E'_s in order to emphasise that the driving quantity E_s and its response E'_s are related to one and the same port. The relevant network equations can then be written as:

$$E'_s = \rho E_s + \nu E_c \tag{1.10}$$

$$E_i = \zeta E_s + \beta E_c \tag{1.11}$$

With the additional relation

$$E_c = AE_i \tag{1.12}$$

and when the relation to be found between E_s and E'_s is an impedance Z_t according to

$$E'_s = Z_t E_s \tag{1.13}$$

we find the following expression for the impedance function:

$$Z_t = \rho \frac{1 - A(\beta - \frac{\nu\zeta}{\rho})}{1 - A\beta} \tag{1.14}$$

Defining now:

$$\beta_o = \left(\frac{E_i}{E_c}\right)_{E_s=0} \tag{1.15}$$

$$\beta_{sc} = \left(\frac{E_i}{E_c}\right)_{E'_s=0} \tag{1.16}$$

the impedance function (1.14) can he rewritten as

$$Z_t = \rho\frac{1 - A\beta_{sc}}{1 - A\beta_o} \tag{1.17}$$

which is Blackman's [9] formula.

Two loop gains $A\beta_{sc}$ and $A\beta_0$ may play a role in the determination of an input or output impedance. In amplifiers with a single feedback loop (to be dealt with in the next section) one of both loop gains is zero while the other is large so that the impedances tend either to zero or infinity. We have seen before that there is no reason to calculate their values because their effects on the amplifier transfer are included in the expression (1.8) for A_t.

In order to realize an accurate impedance, both loop gains have to be large which, as we shall see in the next section, requires at least two feedback loops.

In the ideal case of infinite loop gains (or, in other words, when the active part is assumed to have nullor properties), we find that the impedance obtains a value

$$Z_{t\infty} = \lim_{\substack{A\beta_0\to\infty \\ A\beta_{\sigma\chi}\to\infty}} Z_t = \rho\frac{\beta_{sc}}{\beta_o} \tag{1.18}$$

Expressions (1.17) and (1.18) can be combined to set up an expression in which the impedance is the product of its asymptotic value $Z_{t\infty}$ and of two factors accounting for the non-ideality of the active part, the same as in the asymptotic-gain model. The same loop gains occur, of course, as in Blackman's formula (1.17):

$$Z_t = Z_{t\infty}\frac{-A\beta_o}{1 - A\beta_0}\cdot\frac{1 - A\beta_{sc}}{-A\beta_{sc}} \tag{1.19}$$

This expression can be simplified for those cases where the impedance ρ — the impedance in the case where $A = 0$ — strongly differs from the desired impedance $Z_{t\infty}$. This will usually be true for amplifiers with two feedback loops. In the case where $|\rho| \ll |Z_{t\infty}|$, we find:

$$Z_t = Z_{t\infty}\frac{-A\beta_{sc}}{1 - A\beta_{sc}} \tag{1.20}$$

In the case where $|\rho| \gg |Z_{t\infty}|$, we find:

$$Z_t = Z_{t\infty}\frac{1 - A\beta_{sc}}{-A\beta_{sc}}$$

which can alternatively and for the sake of more uniformity be written as:

$$Y_t = Y_{t\infty}\frac{-A\beta_{sc}}{1 - A\beta_{sc}} \tag{1.21}$$

Expressions (1.20) and (1.21) have the same form as the expression for the gain. However, different loop gains determine the deviations from the intended ideal values of gain and impedance (or admittance). For design purposes it is useful to establish the relation between the two loop gains, which will be done below.

(iii) Relations between transfer and impedance functions

The loop gain $A\beta$ of the amplifier transfer function has to be calculated with the amplifier ports terminated in the source and load impedances. When we indicate the external impedance as Z, we can, according to Blackman [9], write:

$$A\beta = \frac{\rho}{\rho + Z}A\beta_{sc} + \frac{Z}{\rho + Z}A\beta_o \tag{1.22}$$

Using expression (1.18) we can find the desired relations

$$A\beta = \frac{1}{\rho + Z}(Z_{t\infty} + Z)A\beta_o \tag{1.23}$$

$$A\beta = \frac{\rho}{\rho + Z}\left(1 + \frac{Z}{Z_{t\infty}}\right)A\beta_{sc} \tag{1.24}$$

Equation (1.23) is useful in the case where $|\rho| \ll |Z_{t\infty}|$ while equation (1.24) is more suitable for situations where $|\rho| \gg |Z_{t\infty}|$.

These equations can be simplified when Z and $Z_{t\infty}$ have the same order of magnitude, which is the case with characteristic impedance terminations. We then find:

$$A\beta = \left(1 + \frac{Z_{t\infty}}{Z}\right)A\beta_o \tag{1.25}$$

$$A\beta = \left(1 + \frac{Z}{Z_{t\infty}}\right)A\beta_{sc} \tag{1.26}$$

These equations can be used in many situations for a comparison between the impedance and transfer behaviour, in particular of amplifiers with two feedback loops.

Summarising, we can conclude that it is possible to calculate input and output impedances in a way similar to that for the amplifier gain. Clearly, we can determine the ideal transfer function $A_{t\infty}$ together with the ideal impedances $Z_{t\infty}$ as if the active part were a perfect nullor. *The first design step, which involves the determination of the basic amplifier configuration* is thus considerably simplified.

In the subsequent part of this chapter we will make an inventory of basic amplifier configurations, assuming the active part to have nullor properties. Design aspects regarding the loop transfer function(s) $A\beta$ will be put off until later chapters.

1.4 The realisation of transmittances with passive feedback networks

1.4.1 Introduction

The problem of realising the various transmittances as given in table 1.1 might be approached as a mere network-theoretical structural synthesis problem. Such an approach, however, easily loses sight of the practical electronic realisation. We will therefore keep in mind that the active part of a negative-feedback amplifier cannot really be a nullor, but that it produces noise and distortion, that it dissipates power, that it has no perfectly floating ports and that its parameters are frequency dependent. Nevertheless, we will, as far as its transfer properties are concerned, attribute nullor properties to the active part in this and the following sections of this chapter, in order to find the basic amplifier configurations. These configurations will thus be described with the asymptotic gain as defined in section 1.3. In the subsequent chapters we will take full account of the imperfections in the active part.

The elements of the feedback network will be assumed to be ideal and for the time being we will admit all passive network elements, namely the resistance, capacitance, inductance, transformer and gyrator. Though the practical significance of some configurations may be small, this approach is considered to give the best fundamental insight into the electronic realisation problem.

In the next section we will present realisations of all types of transmittances, following a mainly network-theoretical approach. In later sections we will turn our attention to practical realisations, making use of the insight we can gain from general theory. First, we will give some definitions and descriptions that will be useful for our classification purposes.

(i) Some definitions

In this work we will frequently use the *transmission parameters* to describe the transfer properties of a twoport. The twoport equations express the input quantities as functions of the output quantities in the following way:

$$\begin{aligned} U_i &= AU_o + BI_o \\ I_i &= CU_o + DI_o \end{aligned} \quad \text{or} \quad \begin{pmatrix} U_i \\ I_i \end{pmatrix} = \begin{pmatrix} A & B \\ C & D \end{pmatrix} \begin{pmatrix} U_o \\ I_o \end{pmatrix}$$

Signs are attributed to the quantities as indicated in figure 1.5.

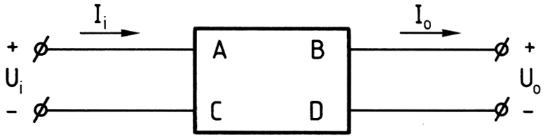


Figure 1.5 Sign conventions for twoports described by their transmission parameters.

Somewhat more familiar to the electronic designer are the reciprocal values of the transmission parameters, which will be referred to as the *transfer parameters*. They are defined as follows:

Voltage-gain factor $\mu = \frac{1}{A} = \left(\frac{U_o}{U_i}\right)_{I_o=0}$

Transadmittance $\mu = \frac{1}{B} = \left(\frac{I_o}{U_i}\right)_{U_o=0}$

Transimpedance $\mu = \frac{1}{C} = \left(\frac{U_o}{I_i}\right)_{I_o=0}$

Current-gain factor $\mu = \frac{1}{D} = \left(\frac{I_o}{I_i}\right)_{U_o=0}$

These definitions implicitly show how to measure the transfer parameters and how to find, indirectly, the transmission parameters.

The active part of the amplifier will — in accordance with the asymptotic gain model — be modelled as a nullor. All transmission parameters of a nullor are zero; all transfer parameters are infinite.

(ii) Description of feedback techniques

In order to find the various basic amplifier configurations with fundamentally different transfer parameters, it is necessary to accurately describe the techniques of applying negative feedback. We will do this first.

The application of negative feedback requires that the output quantity that has to be supplied to the load be sensed by the input port of the feedback network.

- If, for obtaining optimum signal transfer, the load has to be driven from a voltage source, then the voltage across the load must be sensed. This type of feedback is generally called *output shunt feedback*.

- Current sensing, commonly designated as *output series feedback,* is used when the amplifier has to behave as a current source.
- A combination of series and shunt feedback at the output is required when the output power has to be sensed, for example, in order to accomplish characteristic impedance matching.

The sensed output quantity has to be converted accurately in the feedback network into a quantity corresponding to the source quantity that represents the signal information. A comparison is made by summing or subtracting these two quantities at the input of the active part of the amplifier.

- When these quantities are currents, the type of feedback is called *input shunt feedback*.
- In the case of voltages the term *input series feedback* is used.
- In order to realize, for example, characteristic impedance matching a combination of series and shunt feedback at the input is needed.

1.4.2 Classification of non-energic feedback amplifiers

The general criteria formulated in section 1.2 imply the desirability of such a realisation of the feedback action, so that no shunt or series impedances in the signal path are needed. In practice, this cannot be accomplished perfectly, but theoretically, the required feedback networks are available in the form of the ideal transformer and the ideal gyrator. A transformer provides the possibility of converting a voltage into a voltage and a current into a current. The gyrator converts a voltage into a current and vice versa. The gyrator and the transformer are both *non-energic* [10] network elements. No instantaneous power absorption occurs. Noise performance and power efficiency are not degraded when only these ideal elements are used in the feedback network. This was already pointed out by Norton [11] for some transformer-feedback circuits.

Figure 1.6 gives the symbols of the non-energic network elements, together with their transmission matrices.

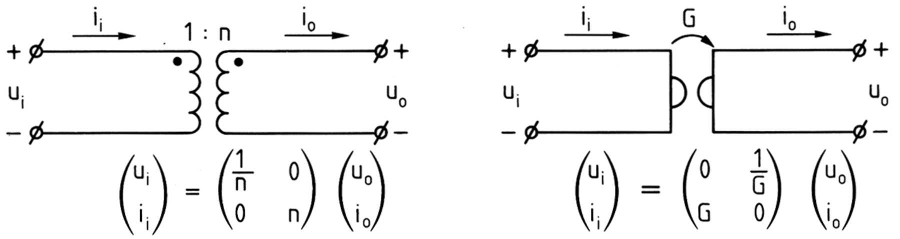

i_i
1 : n
i_o
u_i
u_o
$\begin{pmatrix} u_i \\ i_i \end{pmatrix} = \begin{pmatrix} \frac{1}{n} & 0 \\ 0 & n \end{pmatrix} \begin{pmatrix} u_o \\ i_o \end{pmatrix}$
i_i
G
i_o
u_i
u_o
$\begin{pmatrix} u_i \\ i_i \end{pmatrix} = \begin{pmatrix} 0 & \frac{1}{G} \\ G & 0 \end{pmatrix} \begin{pmatrix} u_o \\ i_o \end{pmatrix}$

Figure 1.6 *Network symbols and transmission matrices of the ideal transformer and gyrator.*

To restrict the number of possible basic-amplifier realisations, the input and output ports of the active part will be given one common terminal, which is also in common with the source and load. Because the source, load and active part have one-side grounded ports, the ports of the feedback elements must be able to float in order to accomplish current sensing and voltage comparison (series feedback at output and input, respectively).

Four feedback loops can be applied to the active part of an amplifier, as indicated in figure 1.7. The feedback elements accomplish the following conversions:

Transformer n_1	Voltage-to-voltage,
Gyrator G_1	Current-to-voltage,
Gyrator G_2	Voltage-to-current,
Transformer n_2	Current-to-current.

We will assume that loop gain is infinite. Consequently, the active part has nullor properties. Its transfer parameters are assumed to have negative signs.

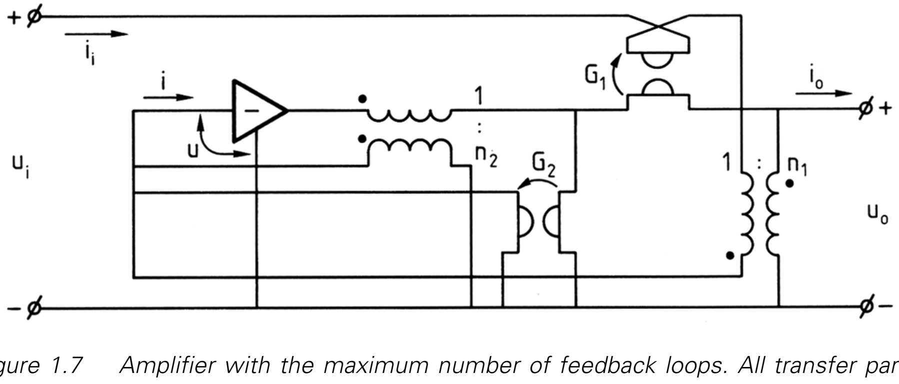


Figure 1.7 *Amplifier with the maximum number of feedback loops. All transfer parameters are determined by the feedback-network transfer function.*

The order of the various sensings is chosen arbitrarily. The exact values of the transfer parameters for this specific order are:

$$\mu = -n_1 \left\{\frac{1 + G_2/G_1 + 1/n_1 n_2}{1 + 1/n_1 n_2}\right\}$$

$$\zeta = -\frac{1}{G_2}\{1 + G_2/G_1 + 1/n_1 n_2\}$$

$$\gamma = -G_1 \left\{\frac{1 + G_2/G_1 + 1/n_1 n_2}{1 + G_2/G_1}\right\}$$

$$\alpha = -n_2\{1 + G_2/G_1 + 1/n_1 n_2\}$$

If all these transfer parameters are large, they are approximately independently determined by the feedback elements, because in that case the direct transfer of current and voltage to the output is small with respect to the transfer via the active part.

By eliminating one or more gyrators or transformers from figure 1.7, less complicated feedback configurations having three loops or less are obtained. Sixteen possible configurations result. Table 1.2 list their *approximate* transfer parameters, assuming that $n_1 n_2 \gg 1$ and $G_1/G_2 \gg 1$.

When these approximations are not allowed, the *exact* expressions of the transfer parameters can easily be found from the above exact expressions for the four-loop amplifier.

Table 1.2. Transfer parameters of the nullor and non-energic feedback configurations with up to four feedback loops (approximate values are given for configurations 6 – 16).

	$-\frac{1}{A} = -\mu$	$-\frac{1}{B} = \gamma$	$-\frac{1}{C} = -\zeta$	$-\frac{1}{D} = -\alpha$	
1	∞	∞	∞	∞	nullor
2	n_1	∞	∞	∞	one loop
3	∞	G_1	∞	∞	
4	∞	∞	$1/G_2$	∞	
5	∞	∞	∞	n_2	
6	n_1	G_1	∞	∞	two loops
7	n_1	∞	$1/G_2$	∞	
8	n_1	∞	∞	n_2	
9	∞	G_1	$1/G_2$	∞	
10	∞	G_1	∞	n_2	
11	∞	∞	$1/G_2$	n_2	
12	∞	G_1	$1/G_2$	n_2	three loops
13	n_1	∞	$1/G_2$	n_2	
14	n_1	G_1	∞	n_2	
15	n_1	G_1	$1/G_2$	∞	
16	n_1	G_1	$1/G_2$	n_2	four loops

Note that each feedback loop essentially fixes the value of one transfer parameter. The properties of the amplifiers will be studied in some detail by inserting them between a signal source and a load as shown in figures 1.8a and 1.8b, where the signal sources are given as voltage and current sources, respectively.

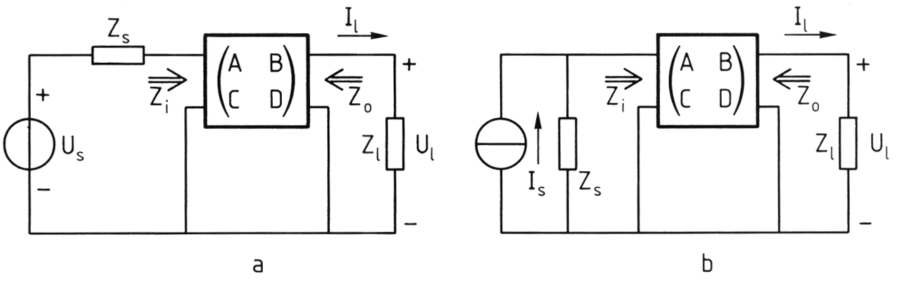


Figure 1.8 Amplifiers inserted between signal sources and loads.

The voltage transfer function of figure 1.8a is given by:

$$\frac{U_\ell}{U_s} = \frac{Z_\ell}{AZ_\ell + B + CZ_s + DZ_s} = A_u \tag{1.27}$$

where A, B, C and D are the transmission parameters of the amplifier. The other transfer functions are related to A_u as:

$$\backslash\mathrm{f}(I_\ell,U_s) = \frac{A_u}{Z_\ell} \qquad \text{(figure 1.8a)}$$

$$\backslash\mathrm{f}(U_\ell,I_s) = A_u Z_s \qquad \text{(figure 1.8b)}$$

$$\backslash\mathrm{f}(I_\ell,I_s) = A_u \frac{Z_s}{Z_\ell} \qquad \text{(figure 1.8b)}$$

The input impedance Z_i and the output impedance Z_o are given respectively by:

$$Z_i = \frac{AZ_\ell + B}{CZ_\ell + D} \quad \text{and} \quad Z_o = \frac{B + DZ_s}{A + CZ_s}$$

The specific properties with respect to the information-transfer aspects of the various configurations will be discussed in the following.

(i) The nullor

Without external circuitry, the nullor is obviously not suitable for the transfer of signal information.

(ii) Single-loop configurations (4)

The four single-loop amplifiers have either zero or infinite input and output impedances and a single well-determined transmission parameter. They can be used when the signal information is represented by a voltage or a current and when the load has to be driven from a voltage or a current source. The source and load impedances have no influence on the transfer functions. Signal-to-noise ratio (SNR) and power efficiency (PE) are exactly equal to those of the active part of the amplifier. This may be elucidated by an example.

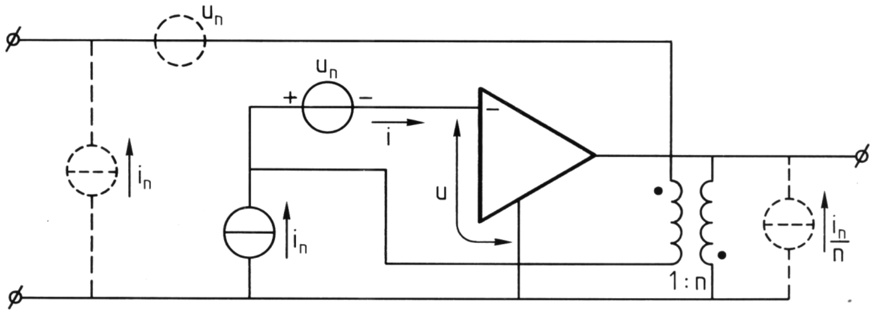


Figure 1.9 *Voltage amplifier with input noise sources of the active part and transformed noise sources indicated by dashes.*

Figure 1.9 shows a voltage-amplifier configuration, with a zero output impedance, an infinite input impedance and a transfer parameter $\mu = -1/n$. All other transfer parameters are infinite. The equivalent input noise sources u_n and i_n of the active part are given in the figure. The result of a transformation of these noise sources is given in the same figure. The transformation techniques used are explained in section 3.2.

The source i_n/n does not contribute to the equivalent input noise sources, because it is transformed into input sources by dividing it by the transfer parameters α and γ, which are both infinite. The equivalent input noise sources are consequently equal to those of the active part. Furthermore, there is no power loss at the output, because the current in the transformer equals zero. Similar considerations can be given for the other single-loop configurations.

(iii) Dual-loop configurations (6)

The combination of shunt and series feedback either at the input or at the output or at both input and output leads to the possibility of realising accurate power transfer which is needed for characteristic impedance matching.

The input impedance may or may not depend on the load impedance. If it does, the amplifier is not *unilateral,* and as a consequence, accurate matching at the input port is possible only if the load impedance is accurately known. If it does not, the matching at the input port is independent of the situation at the other port, where the driving point impedance is either zero or infinite. The output impedance exhibits a similar behaviour, in relation to the source impedance.

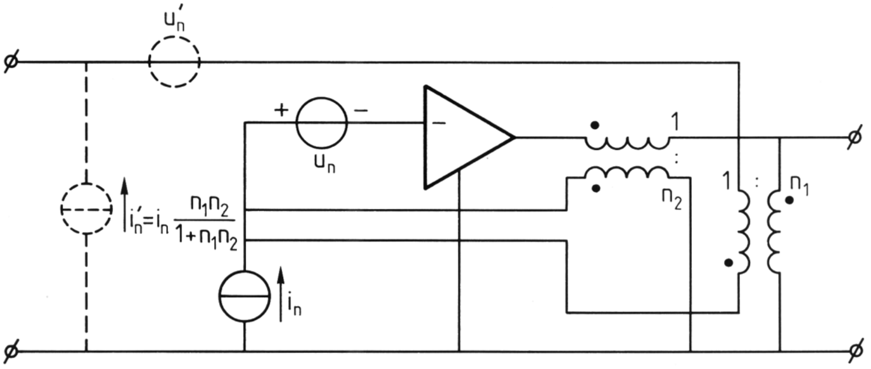


Figure 1.10 Transformation of the noise sources of the active part to the input of a dual-loop amplifier.

In some cases, the feedback networks cause some improvement of the SNR compared to that of the active part. An example is given in figure 1.10, where the equivalent input noise sources of the active part are transformed into input noise sources of the amplifier. The noise current source i'_n is somewhat smaller than i_n, while the sources u'_n and u_n are- equal. The resulting SNR improvement (too small to be useful) is a consequence of the increase in available output power by the direct transfer, via the feedback network, from source to load. This relative share of the directly transferred source power in the available output power gets smaller as the available power gain gets larger. The impedance and transfer properties of all possible (6) dual-loop feedback amplifier types are summarised in table 1.3. The configuration numbers refer to table 1.2. Configurations 8 and 9 can be used for characteristic impedance matching at input and output ($Z_i = Z_s$, $Z_o = Z_\ell$), because $Z_iZ_o = Z_\ell Z_s$ and $Z_iZ_\ell = Z_oZ_s$, respectively. For equal impedance levels at input and output, A and D must be equal in configuration 8, while in configuration 9 the condition that $B/C = Z_sZ_\ell$ must be met. Matching errors at one port are reflected at the other port. As a consequence, these configurations are not ideal for matching purposes.

(iv) Three-loop configurations (4)

Though it is possible to realize accurate input and output impedances with these configurations when load and source impedances are accurately known, there is no possibility for characteristic matching at both sides. The interdependence of Z_i and Z_ℓ and of Z_o and Z_s cannot be avoided. The configurations are therefore not attractive for any application.

Table 1.3 Transfer properties of dual-loop feedback amplifiers.

					Z_i	Z_o	transfer function
6	A	B			∞	$\frac{B}{A}$	$\frac{U_\ell}{U_s}=\frac{Z_\ell}{AZ_\ell+B}$
7	A		C		$\frac{A}{C}$	0	$\frac{U_\ell}{I_s}=\frac{Z_s}{A+CZ_s}$
8	A			D	$\frac{AZ_\ell}{D}$	$\frac{DZ_s}{A}$	$\frac{U_\ell}{U_s}=\frac{Z_\ell}{AZ_\ell+DZ_s}$
9		B	C		$\frac{B}{CZ_\ell}$	$\frac{B}{CZ_s}$	$\frac{U_\ell}{U_s}=\frac{Z_\ell}{B+CZ_s/Z}$
10		B		D	$\frac{B}{D}$	∞	$\frac{I_\ell}{U_s}=\frac{1}{B+DZ_s}$
11			C	D	0	$\frac{D}{C}$	$\frac{U_\ell}{I_s}=\frac{Z_\ell}{CZ_\ell+D}$

(v) Four-loop configuration (1)

All parameters are determined accurately in the case of four feedback loops. The parameters can be designed so that input and output impedances do not depend on source and load impedances. The condition for this independence is:

$$AD = BC, \quad \text{or} \quad \zeta\gamma = \mu\alpha$$

In that case:

$$Z_i = A/C = B/D \quad \text{and} \quad Z_o = B/A = D/C$$

Impedance matchings at input and output are not interdependent. With characteristic matching at both sides, the voltage transfer function is given by:

$$\frac{U_\ell}{U_s} = \frac{1}{4A}$$

(vi) Concluding remarks

The configurations which were presented in this section are paradigms of negative-feedback amplifiers, where sensing of the output quantities and comparison of the input quantities are accomplished in an ideal, non-energic way. Moreover, each feedback loop virtually independently determines one transfer parameter provided that the direct power transfer from source to load is relatively small. The theoretical significance of this classification is great because the configurations form a complete set of ideal amplifier types. Their practical significance is slight because the ideal gyrator cannot be approximated well enough.

As soon as impedances are used in the feedback networks, the ideal situation is lost. For example, current sensing will cause a voltage drop and voltage sensing will cause extra current flow. Noise may be generated in the feedback impedances and SNR and PE may be deteriorated. The class of non-energic feedback amplifiers admits realisations of transfer functions with either positive or negative signs, thanks to the floating ports of the feedback networks.

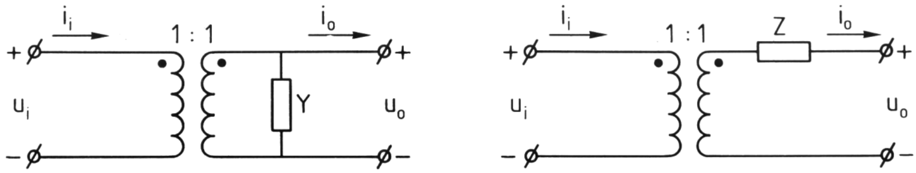
Practical amplifiers must necessarily have feedback networks with floating ports too, in order to obtain transfer properties similar to the configurations discussed. If active feedback networks are not considered, these practical realisations must use transformers for obtaining inverting as well as non inverting alternatives for the gyrator feedback. A small number of transfer function types can be realised without transformers. The class of transformer feedback configurations is presented in the next section.

1.4.3 Classification of amplifier configurations with transformer feedback

A physical realisation of the transformer can approximate the ideal transformer to a large extent, provided that the impedance and power levels at which it is used are low, that its physical dimensions are small (low stray capacitance) and that frequencies are not too low. The transformer (e.g. wound on a toroid core) can then be modelled over a number of frequency decades as an ideal transformer with floating ports.

The gyrator is much more difficult to realize in a useful physical form. Electronic realisations may approximate the ideal in a few low-frequency decades, but only as far as non-energicness is concerned. Their noise production and non-linearity — inherent when active devices and resistances that fix the gyration constants are used — render them useless as feedback elements in practical wide-band amplifiers.

The gyrator function in the feedback network can be performed, however, by impedances in combination with transformers. In some cases the transformers can be omitted. Figure 1.11 shows the alternatives for the gyrator with the corresponding transmission matrices. For the sake of simplicity the transformers will have unity turns ratios. The active part of the amplifier will have one-side-grounded ports again and will invert the input quantities (μ, γ, ζ and α all have values $-\infty$). In some special cases we will forego this restriction.



$$\begin{pmatrix} u_i \\ i_i \end{pmatrix} = \begin{pmatrix} 1 & 0 \\ Y & 1 \end{pmatrix} \begin{pmatrix} u_o \\ i_o \end{pmatrix} \qquad \begin{pmatrix} u_i \\ i_i \end{pmatrix} = \begin{pmatrix} 1 & Z \\ 0 & 1 \end{pmatrix} \begin{pmatrix} u_o \\ i_o \end{pmatrix}$$

Figure 1.11 Transformer-immittance combinations to be used as practical alternatives for the gyrator.

The one-sided grounding of the ports of the active part has a practical significance now. Perfectly floating ports are hard to realize. An infinite value of the common-mode rejection ratio is needed at the input, and perfectly equal currents have to flow in the output leads. Parasitic effects result if these conditions are not met. Input and output impedances are shunted in that case by parasitic impedances, and distortion may be enhanced. We will take these imperfections into account in chapter 4. Figure 1.12 depicts the configuration with four feedback loops. These are not completely independent now. Current sensing, for example, influences the voltage sensing and vice versa.

The following equations hold if nullor properties are attributed again to the active part:

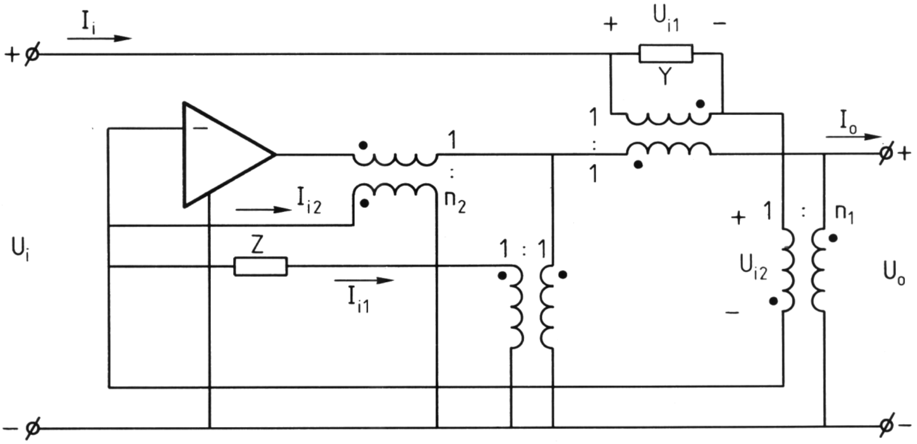


Figure 1.12 General feedback configuration with four feedback loops implemented with transformers and immittances.

$$U_i = U_{i1} + U_{i2}, \quad I_i = I_{i1} + I_{i2}$$

$$-I_{i1}\left(\frac{n_1 - 1}{n_1}\right) + I_{i2}\left(\frac{n_1 n_2 + 1}{n_1}\right) = -I_o$$

$$U_o - U_{i1} = -I_{i1}Z, \quad U_o = -n_1 U_{i2}, \quad U_{i1} = \frac{I_i}{Y}\left(\frac{n_1 - 1}{n_1}\right) - \frac{I_o}{Y}$$

All parameters of all 16 possible configurations can be determined from these equations. We will discuss their properties briefly below.

(i) The active part of the amplifier

The active part of the amplifier will generally not be suitable for information transfer unless additional circuitry is used. It produces power gain in order to realize the desired magnitude of the transfer function as well as to obtain a high quality of information transfer when used in combination with one or more negative-feedback loops.

(ii) Single-loop configurations (figure 1.14)

The four single-loop amplifiers, derived directly from figure 1.12 by making three out of the four quantities Y, Z, n_1 and n_2 infinite, have transfer parameters that are respectively given by:

Voltage amplifier	$\mu = -n_1$,
Transadmittance amplifier	$\gamma = -Y$,
Transimpedance amplifier	$\zeta = -Z$,
Current amplifier	$\alpha = -n_2$.

The transformer can be omitted in the transimpedance amplifier, provided that the active part is inverting. The μ and α realisations need transformers exclusively and are therefore essentially non-energic. Input and output impedances of the configurations are either zero or infinite.

Non-inverting unity voltage-gain and unity current-gain amplifiers can be realised also without the aid of transformers provided that the amplifier has floating ports. The type of feedback is non-energic, but they have the disadvantages of parasitic effects, as mentioned before. Figure 1.13 shows the symbol adopted for an active part with floating ports. Figure 1.14 shows the single-loop configurations with their more or loss currently used names.

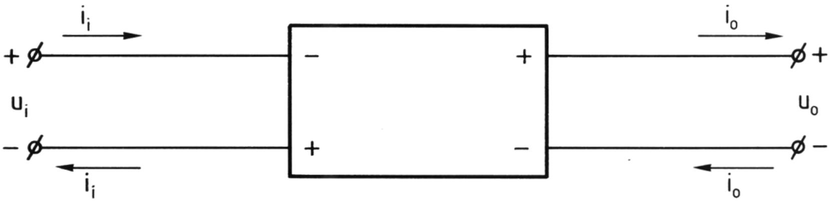


Figure 1.13 Symbol for ideally floating active part.

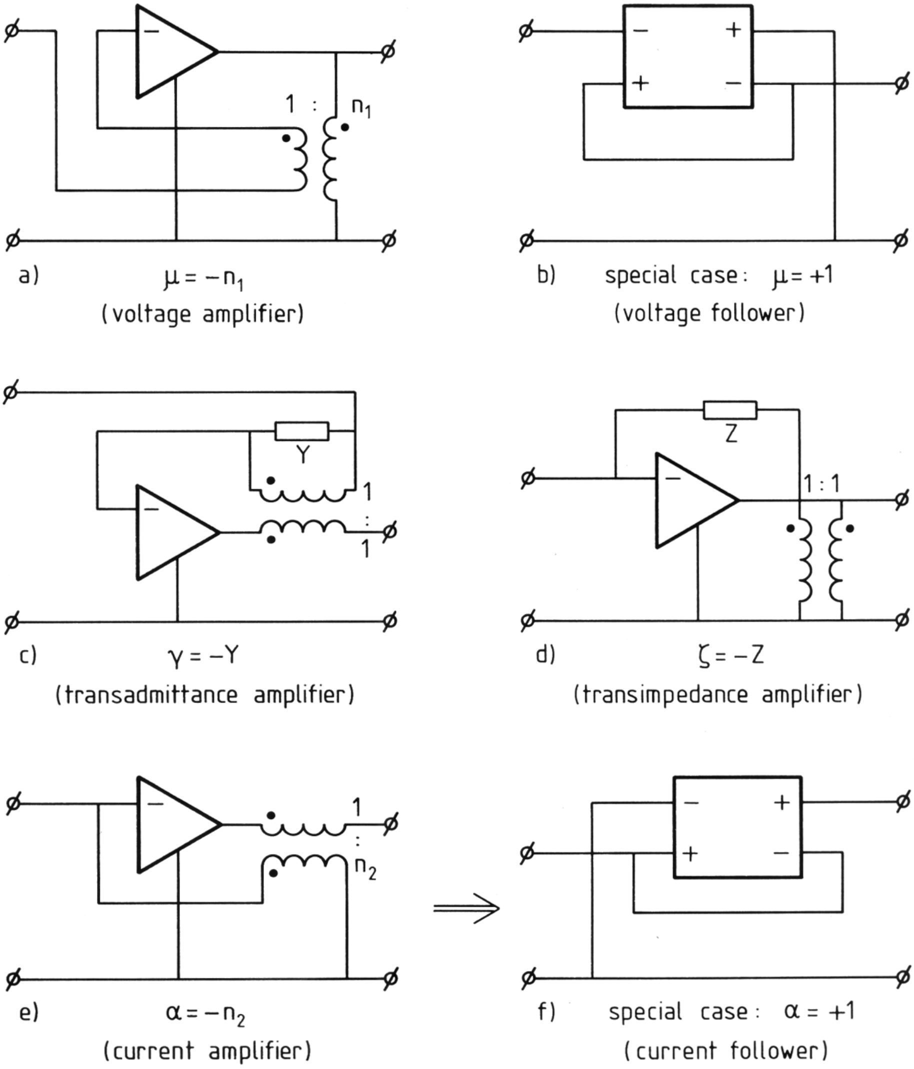


Figure 1.14 Single-loop amplifiers with zero or infinite input and output impedances.

(iii) Dual-loop configurations

The amplifiers with two transformer-feedback loops form a group that — to the author's knowledge — has not been classified as such before. Isolated examples can be found in literature [12], [13], [14].

Non-energicness is violated in all cases where impedances are used. The deterioration of SNR and PE is determined by the types and the magnitudes of the impedances in the feedback networks, together with the turns ratios of the transformers. Large values of Z and Y are advantageous, as will be shown in chapters 3 and 4. The configurations are able to accomplish characteristic impedance matching at input and/or output, just like the non-energic types presented in section 1.4.2.

We will discuss briefly the properties of the configurations that follow directly from figure 1.12 by making two out of the four quantities Y, Z, n_1 and n_2 infinite. Some alternative configurations, based on the voltage and current followers of figure 1.14b and f, are presented in addition.

- The four configurations with either an accurate input or an accurate output impedance are given in figure 1.15. Their transfer properties correspond to configuration numbers 6, 7, 0 and 11 from table 1.3. They are essentially unilateral. For obtaining these specific transfer and impedance properties with passive feedback networks, transformers are indispensable. Only the transimpedance parameters can be fixed without a transformer, provided that the active part is inverting. Non-energicness is lost as a result of the use of impedances in the feedback networks. By cascading the appropriate configurations, characteristic-impedance matching at both input and output can be realised.
- The two non-unilateral configurations with transfer properties corresponding to the configuration numbers 8 and 9 in table 1.3 are given in figure 1.16. Figure 1.16a shows the two-transformer configuration which has essentially non-energic feedback. Figure 1.16b represents an approximation of the two gyrator configuration. Alternative dual-loop non-unilateral configurations with transfer properties similar to the two-transformer configuration of figure 1.16a can be obtained with the follower circuits of figures 1.14b and f in combination with only one transformer. The gain factors of these followers can be raised above unity with the aid of auto-transformers so that a voltage amplifier and a current amplifier result, as shown in figure 1.17a and b, respectively.

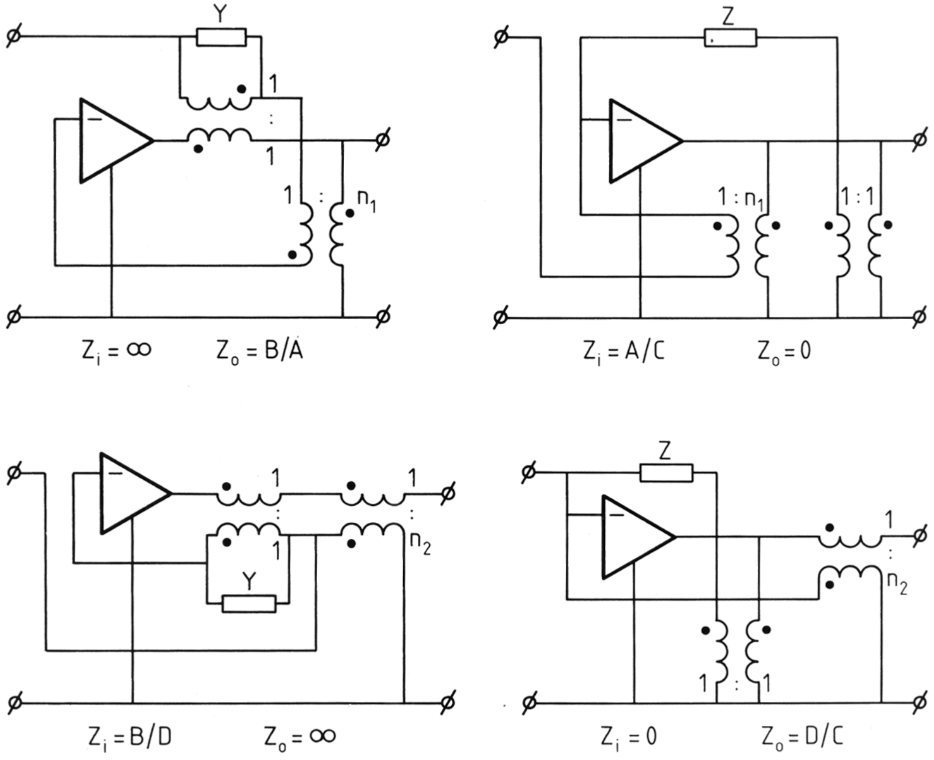


Figure 1.15 *Dual-loop configurations with accurate impedance at one port and either zero or infinite impedance at the opposite port.*

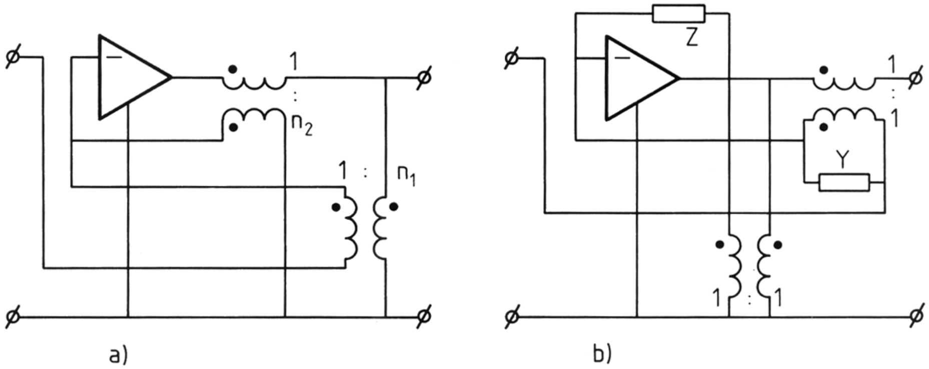


Figure 1.16 *Non-unilateral dual-loop configurations.*

By providing the transformers with secondary windings, the output current in the circuit of figure 1.17a can be sensed and a part of it can be fed back. In figure 1.17b the same can ho done with the output voltage. Figure 18 shows these extensions. The circuit of figure 1.18d is described in a patent by Norton [14], where the active part is formed by a single CB stage. All configurations of figure 1.18 have essentially non-energic feedback, but suffer from the disadvantage of parasitic effects duo to the floating ports of the active part. Moreover, matching errors at one port are reflected at the other port.

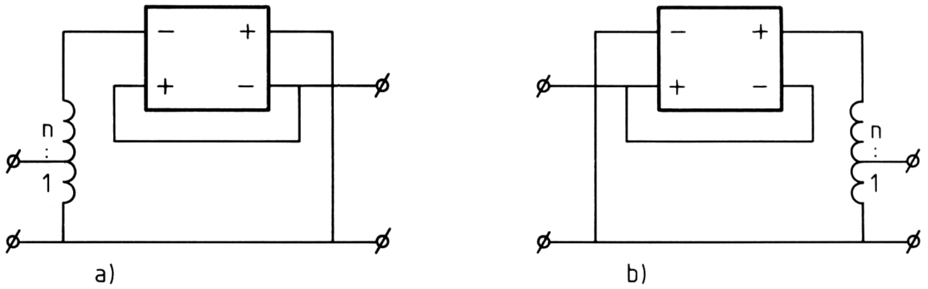


Figure 1.17 Voltage-gain factor μ and current-gain factor α of the follower configurations enlarged to a value n with the aid of auto-transformers.

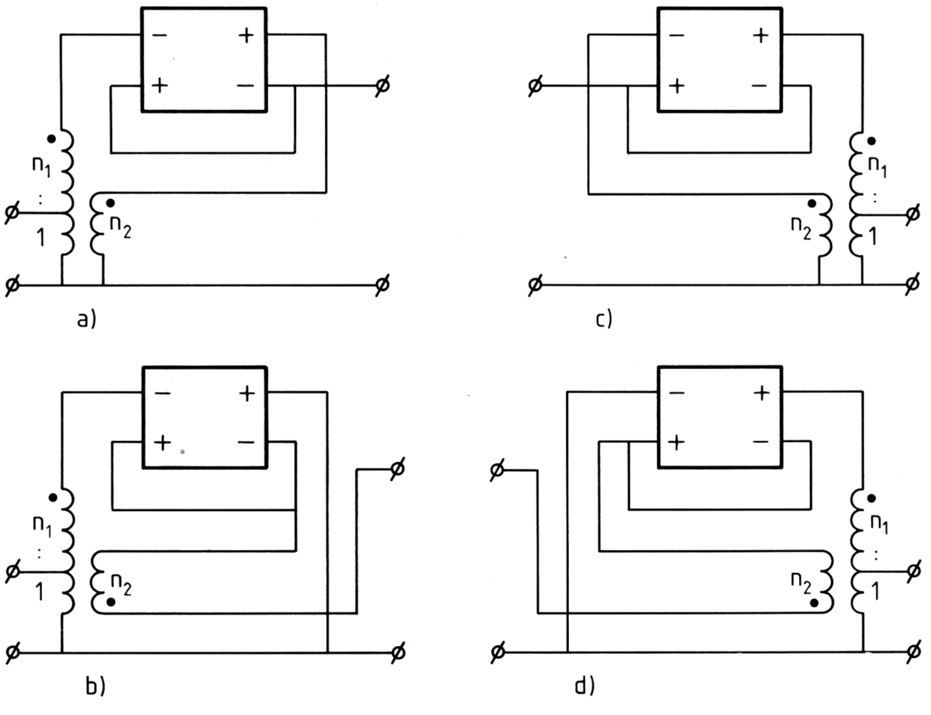


Figure 1.18 Alternative configurations for non-energic transformer feedback, using followers and auto-transformers.

(iv) Three-loop configurations

Realisation of three-loop amplifiers will not be discussed because of their small practical significance.

(v) Four-loop configuration

The parameters of the four-loop configuration are given by fairly complex expressions. To get an impression of the properties of this amplifier, some reasonable approximations will therefore be made, namely: $n_1 \gg 1$, $n_2 \gg 1$, $YZ \gg 1$.
The parameters B and D are determined with a short-circuited output, i.e. $U_o = 0$. With the assumption that $I_i \ll I_o$, we find:

$$B \approx \frac{-1}{Y}, \quad D \approx -(1/n_2 + 1/YZ)$$

The parameters A and C are determined with an open output; i.e. $I_o = 0$.
We find:

$$A \approx -(1/n_1 + 1/YZ) \quad \text{and} \quad C \approx -1/Z$$

A unilateral transfer is possible when the condition $AD = BC$ is met, or when

$$Y^2Z^2 + (n_1 + n_2 - n_1n_2)YZ + n_1n_2 = 0$$

which reduces with the given condition to $YZ = n_1n_2$. We will not work out the possibilities of this rather complicated feedback configuration any further.
We will conclude this section by referring to a circuit having four feedback loops around one active device which was patented in 1971 [15]. By using an active part with floating ports (one transistor), two transformers can be omitted. The circuit mentioned has equal input and output impedances of 50 Ω, quite insensitive to errors in the source and load impedances. The transformer feedback is realised with a directional coupler. Figure 1.19 shows the basic circuit diagram.

1.4.4 Realisation of transmittances without the use of transformers

There are many practical situations where the use of transformers cannot be tolerated, either because of their loss favourable electrical properties, especially at low frequencies, or for economical reasons. We therefore have to look for other realisations of the feedback networks using impedances exclusively. For reasons of interference rejection it is generally required that source, load and amplifier have one common terminal. In that case (the only one to be dealt with here), equipping the active part with floating ports according to figure 1.20 for current sensing at the output and for voltage comparison at the input cannot be avoided.

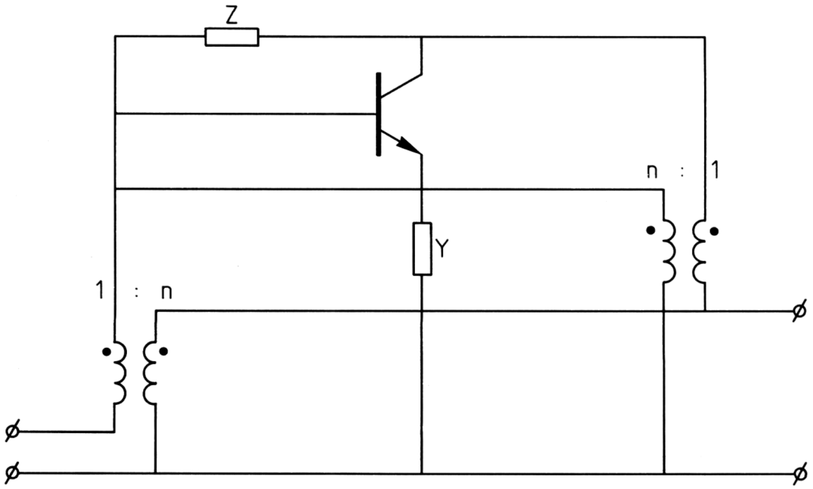


Figure 1.19 Amplifier with four feedback loops using a directional coupler.

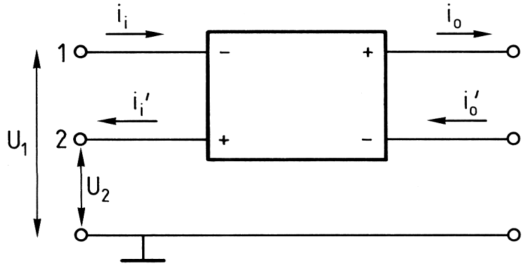


Figure 1.20 Active part with ports floating with respect to ground.

For perfect current sensing, the currents i_o and i'_o (figure 1.20) must be exactly equal. For a perfect comparison of the source voltage and the feedback voltage, the common-mode rejection of the active part must be infinitely large. In other words: the voltage sensitivities of both input terminals must be exactly equal, but must have opposite signs.

The ideal situation cannot, of course, be realised in practice. The active part is connected via finite impedances to a power supply and has finite physical dimensions. Therefore, it is either directly or via parasitic impedances connected to ground, and the common-mode rejection cannot be infinitely large. These imperfections in the active part lead to departures from the ideal asymptotic-gain value. We will ignore them at this stage and account for them in chapter 4.

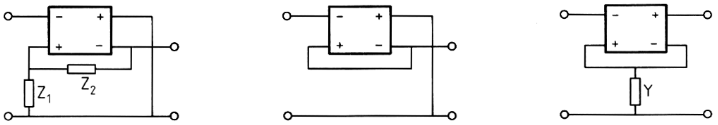
The configurations that can be realised with the aid of active parts with floating ports and passive components, except transformers, are given in figure 1.21. The number of basically different realisations is reduced to only six. Some special cases of the voltage and current amplifier and of the two-loop configurations are added. All other configurations listed in table 1.2, section 1.4.2 require inverting feedback networks, which can either be realised with transformers or with active devices. We dealt with transformers in the preceding section. Active feedback networks will be treated concisely in the next section.

The values of the parameters of the basic configurations are given in the figure captions in figure 1.21. Only the configurations with unity-gain factors are non-energic. The other configurations have impedances in series or in parallel with the signal path, so that signal-to-noise ratio and efficiency are somewhat deteriorated as will be shown in chapters 3 and 4.

Although the series impedances can have small values, their effect on input or output impedances can be large. Shunt impedances may be largo and still have a great effect on impedances as well. In the transadmittance and transimpedance amplifiers, the influences of the series and shunt impedances on the signal-handling capability and signal-to-noise ratio are fixed if the transfer functions must have certain prescribed values. Compromises are possible in the voltage and current amplifiers, where for optimum noise performance Z_1 must be small and Z_2 must be large, respectively. For maximum efficiency, Z_2 must be large and Z_1 must be small. For given values of the voltage or current gain these requirements may be conflicting, so that efficiency must be sacrificed in favour of signal-to-noise ratio, or vice versa.

Lossless impedances can sometimes be used in the feedback paths. Though they produce no noise, they can have an adverse influence on the signal-to-noise ratio and though they dissipate no signal power, signal handling may decrease. In chapters 3 and 4 we will pay more attention to these aspects.

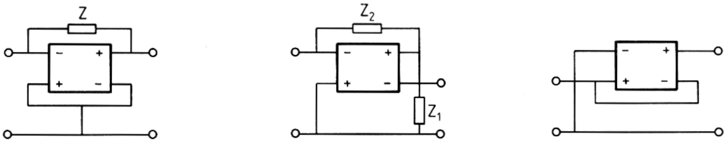
The first six configurations in figure 1.21 have single feedback loops. When loop gain is infinite, the input and output impedances are either zero or infinite. Their transmittances are completely determined by the passive components of the feedback networks. The remaining configurations have two feedback loops and are capable of realising accurate input and output impedances, provided that the source and load impedances are accurately known. The features of the two-loop configurations will be studied in some detail.



a. $\mu = +(1+Z_2/Z_1)$
(voltage amplifier)

b. special case: $\mu = 1$
(voltage follower)

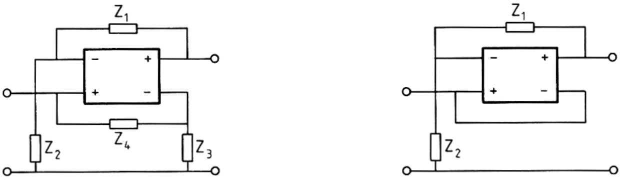
c. $\gamma = -Y$
(transadmittance amplifier)



d. $\zeta = -Z$
(transimpedance amplifier)

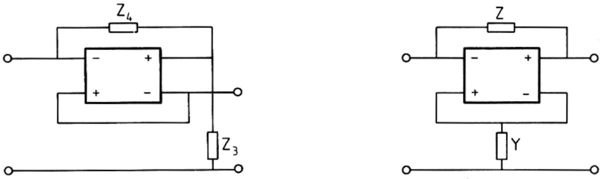
e. $\alpha = +(1+Z_2/Z_1)$
(current amplifier)

f. special case: $\alpha = +1$
(current follower)



g. $\mu = (Z_2+Z_1)/Z_2$, $\gamma = \infty$
$\zeta = (Z_1+Z_2)(Z_3+Z_4)/(Z_2+Z_3)$,
$\alpha = (Z_3+Z_4)/Z_3$

h. special case: $\mu = (Z_1+Z_2)/Z_2$, $\gamma = \infty$
$\zeta = Z_1 + Z_2$, $\alpha = 1$



i. special case: $\mu = 1$, $\gamma = \infty$
$\zeta = Z_3 + Z_4$, $\alpha = (Z_3+Z_4)/Z_3$

j. $\mu = 1 - YZ$, $\gamma = (1-YZ)/Z$
$\zeta = (1-YZ)/Y$, $\alpha = 1 - YZ$

Figure 1.21 Amplifier configurations with passive feedback components with the exception of transformers.

(i) Two-loop configuration of figure 1.21g

Configuration 1.21g has the following transmission parameters:

$$A = \frac{Z_2}{Z_1 + Z_2}, \quad B = 0, \quad C = \frac{Z_2 + Z_3}{(Z_1 + Z_2)(Z_3 + Z_4)}, \quad D = \frac{Z_3}{Z_3 + Z_4}$$

It has the characteristics, therefore, of a three-loop amplifier as discussed in section 1.4.2.

With

$$Z_i = \frac{AZ_\ell}{CZ_\ell + D} \quad \text{and} \quad Z_o = \frac{DZ_s}{A + CZ_s}$$

it follows that an accurate input impedance can be realised, provided that the load impedance Z_ℓ is accurately known, or infinitely large. An accurate output impedance is obtained when the source impedance Z_s is accurately known or infinitely large. Characteristic matching at both the input and the output is not possible in an accurate way, unless a series impedance is inserted in the input or output leads in order to fix parameter B.

The configuration is suitable, for example, for realising an accurate and linear low-noise damping resistance for magneto-dynamic transducers [16]. Moreover, a frequency-dependent transfer function can be realised at the same time [17]. The circuit is frequently used for impedance matching at input and output [18] in spite of its inferiority compared with the circuit of figure 1.21j.

(ii) Two-loop configuration of figure 1.21j

The transmission parameters of the configuration in figure 1.21j are given by:

$$\mathrm{A} = \frac{1}{1 - YZ}, \qquad \mathrm{E} = \frac{Y}{1 - YZ}$$

$$\mathrm{B} = \frac{Z}{1 - YZ}, \qquad \mathrm{D} = \frac{1}{1 - YZ}$$

Though all parameters have accurate values, the configuration does not have the versatile properties of the ideal four-loop amplifier. The condition for realising input and output impedances that are independent of source and load impedances coincides with a zero transfer condition. In other situations, the configuration is not unilateral.

Characteristic impedance matching at both sides is possible under the condition:

$$R_c^2 = \frac{Z}{Y},$$

where $R_c = R_i = R_s = R_o = R_\ell$.

The transfer function can be frequency dependent at the same time if Z and Y have the same frequency dependency. The circuit is, in principle, more suited to realising input and output impedance matching than the configuration of figure 1.21g. Nevertheless, it is probably used less for this purpose [19], [20].

1.5 Realisation of transmittances with active feedback networks

1.5.1 Introduction

In the previous sections we have seen that a number of amplifier types can only be realised with the aid of inverting feedback elements. The only passive feedback element that can be used in practice for this purpose is the transformer. In order to enable the designer to realise, for example, inverting current amplifiers, or amplifiers with an accurate impedance at one port and a zero or infinite impedance at the other, without the use of transformers, we have to resort to active feedback networks that have to accomplish the required inversion.

It will appear to be possible and meaningful to distinguish two fundamentally different methods for realising negative feedback with active feedback networks. One will be referred to in the following as the *active-feedback* technique. It is similar to the passive-feedback technique in so far as the output quantity itself is sensed and the feedback quantity is compared directly with the input quantity. The other, which will be called an *indirect-feedback* technique, does not — when applied to the output — sense the output quantity itself, but instead a quantity which resembles the output quantity as much as possible. An indirect comparison of the input quantity with the feedback quantity is equally possible. The indirect-feedback method thus bears some resemblance to pantographic techniques.

It should be noted that the active- and indirect-feedback techniques are basically inferior to passive-feedback techniques with respect to nearly all quality aspects. The configurations are interesting in the first place because they supplement the number of transmittance realisations. Moreover, some of them are especially attractive for implementation as integrated circuits.

1.5.2 Indirect-feedback amplifiers

The use of an indirect-feedback technique may be considered when a transmittance with a sign opposite to that obtained in the corresponding passive-feedback configuration is needed, or when an accurate input or output impedance at one port of the amplifier is required.

Indirect feedback at the output of an amplifier differs from the direct-feedback methods discussed up to now in so far as the actual quantity supplied to the load is not sensed. Instead of sensing the actual load current or voltage, the output quantity of a dummy output stage is sensed, and the load quantity is supplied by a similar second stage. The transfer function of this actual output stage is not enclosed by the feedback loop. For obtaining an accurate and linear amplifier transfer, certain parameters of both output stages have to be equal. Furthermore, their relevant transfer characteristics have to be linear or equally non-linear.

Indirect-feedback techniques at the input can be described in a similar way. The examples that will be given below are self-evident.

The indirect-feedback techniques to be discussed may not be basically now. They are similar or at least related to techniques described earlier [21], [22]. The explicit formulation and generalisation of these techniques as feedback techniques, however, is believed to be justifiable because it may be useful in various amplifier design problems.

We will not discuss all possible types of indirect-feedback configurations, but will instead confine ourselves to examples of configurations that yield the most interesting transfer properties from the viewpoint of utility.

(i) The technique of indirect current sensing

Figure 1.22 shows, as an example, the principle of indirect current sensing combined with direct current comparison (shunt feedback) at the input. Two equivalent output stages I and II are needed. The output current I_o of stage I instead of the actual output current I_ℓ is sensed, and part of this current is fed back to the amplifier input. The bias conditions in the output stages are assumed to be equal. The transfer properties of the output stages are characterised by their transmission parameters, and the loop gain in the direct-feedback loop (formed by the active part AP, output stage I and the current divider Z_1-Z_2) is assumed to be infinite. In that case U and I are necessarily zero and the asymptotic value of the current gain is given by:

$$A_{i\infty} = \frac{I_\ell}{I_s} = -\left(1 + \frac{Z_2}{Z_1}\right)\left(\frac{AZ' + \mathrm{B}}{A'Z_\ell + B'}\right) \tag{1.28}$$

where $$Z' = \frac{Z_1 Z_2}{Z_1 + Z_2}$$

The parameters C and D of the output stages obviously don't play any role in the transfer function in this ideal situation. Effects of a finite transimpedance ($1/C$) and of the current-gain factor ($1/D$) can therefore be reduced by increasing the loop gain in practical situations.

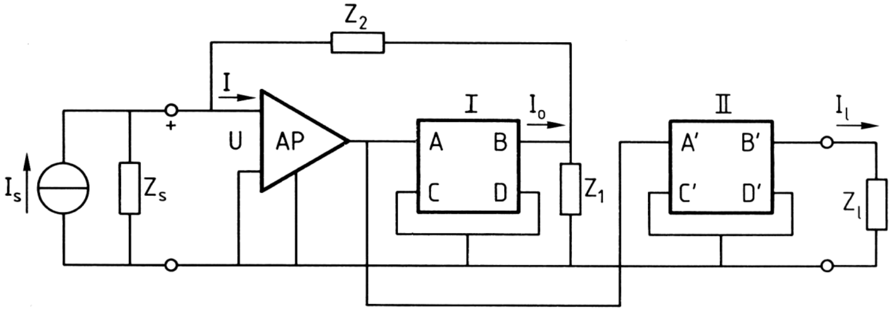


Figure 1.22 Current amplifier with indirect current sensing.

In a direct-feedback amplifier, the stage parameters need not influence the transfer function when the loop gain is infinite. In this indirect-feedback situation, however, output-stage parameters A and B occur in the transfer function of the amplifier, and because their values are signal dependent, linearity is not guaranteed even when the loop gain is infinite. Special measures for the reduction of this non-linearity may be necessary. The ratio of the parameters A and B represents the output impedance of a stage under voltage-drive conditions. The non-linearity can therefore alternatively be ascribed to the signal-induced variations of this output impedance.

For obtaining perfect linearity, it is necessary that signal voltages and currents in both stages be equal. By meeting the condition that $Z' = Z_\ell$ follows from (1.28) that the transfer function is determined by the passive impedances Z_1 and Z_2 exclusively because, in that case, A and A' as well as B and B' can be mutually equal, resulting in a perfect compensation. An obvious disadvantage of meeting this condition, however, is that both output stages have to handle equal signal powers and, in fact, an output-power loss of 3 dB results. Moreover, the value of Z_ℓ must be well known.

The power loss can be reduced by using several identical configurations in parallel in stage II. When n configurations are paralleled in stage II, the parameter A' remains unaltered, whereas the parameter B' is reduced by a factor n, provided that the bias conditions in all configurations are equal. When the output stages can be implemented so that the influence of the terms with A and A' can be disregarded, the current gain of this amplifier becomes:

$$A_{i\infty} = \frac{I_\ell}{I_s} = -n(1 + Z_2/Z_1)$$

This current transfer is linear provided that the transconductances of all output configurations vary in the same way along with the signal. The signal power handled by output stage I can thus be significantly smaller than that handled by output stage II. Large amounts of output power loss are avoided.

Scaling the gain in accordance with the number of paralleled output stages offers the possibility of realising indirect-feedback configurations with arbitrary current gains without using any passive impedances. A very simple example is given in figure 1.23, where CE-configurations are used as output stages. The configuration with an inverting *unity current gain* is generally known as a *current mirror*.

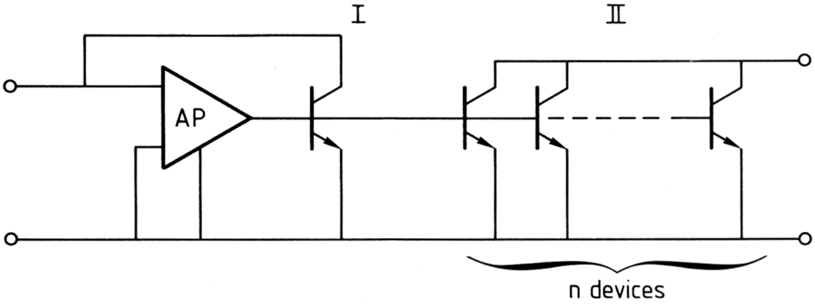


Figure 1.23 Current amplifier with gain scaling achieved by using n paralleled devices in stage II.

(ii) Output-stage implementation

The question arises, as a matter of course, of how the output stages of figure 1.22 can be implemented so that the influence of transmission parameter A can be disregarded ($AZ_\ell \ll B$). It will be clear that we have to be content with approximations that will be valid only in cases where the load impedances are relatively small. A single transistor in a common-emitter or common-source configuration, as in figure 1.23, may be a reasonable choice when the load impedance is low, but more elaborate output stages, whether or not with negative feedback, may give much better results.

At this stage it is appropriate to pay some attention to the unique capabilities of common-emitter differential output stages. Figure 1.24 shows an indirect-feedback current amplifier equipped with such stages. When biased at equal tail currents, the circuit behaves like a unity-gain current amplifier with both an inverting and a non-inverting output. Thanks to the exponential relationship between base-to-emitter voltage and collector current in *a bipolar* transistor (or in an MOS transistor in the subthreshold region), and consequently, to the fact that its transconductance is linearly proportional to the collector current, the current gain of this amplifier can be varied by varying the ratio of the tail currents. For small load impedances ($AZ_\ell \ll B$), the current transfer characteristic remains linear, independent of this ratio. The circuit of figure 1.24 can alternatively he considered to operate according to the *translinear principle,* as first formulated by Gilbert [23]. Such a point of view, however, fits less well in a classification of possible negative-feedback configurations.

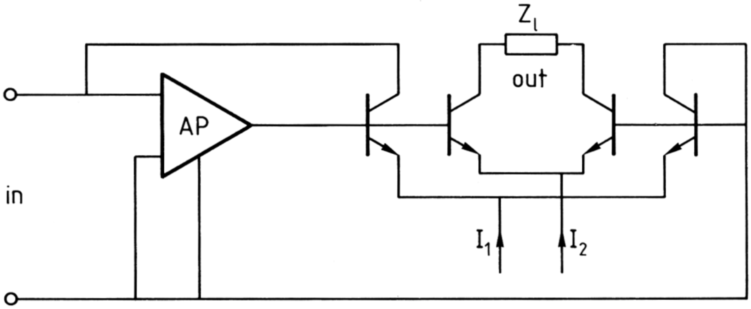


Figure 1.24 Indirect–feedback current amplifier with differential output configurations.

Further aspects of output-stage implementation will be put off until later chapters, where the influence of the output stage parameters on transfer linearity and accuracy will be studied in more detail.

(iii) Indirect current sensing combined with input series feedback

Instead of realising low input impedances with the aid of input shunt feedback, as in the configurations discussed up to this point, it is equally possible to realise high input impedances by means of input series feedback. A basic configuration of such a transadmittance amplifier is shown in figure 1.25. The active part must have a floating input port, just as in conventional transadmittance amplifiers. Assuming an ideal active part, the asymptotic value of the transadmittance is given by:

$$Y_\infty = \frac{I_\ell}{U_s} = Y\left(\frac{A/Y + B}{A'Z_\ell + B'}\right) \tag{1.29}$$

The factor in parentheses has the same form as in expression (1.28), and the discussions in connection with that expression apply equally well to this configuration.

A noticeable feature of an indirect-feedback transadmittance amplifier is that it combines a high input impedance with the possibility of obtaining a variable gain by using differential output configurations similar to those of the indirect-feedback current amplifier of figure 1.24. This opens up the possibility of using an additional passive feedback loop in order to realise an accurate and linear input impedance that can he controlled by varying the ratio of the currents in the output stages. Such a current-controlled impedance may he useful, for example, for automatic gain control in oscillators or in filters.

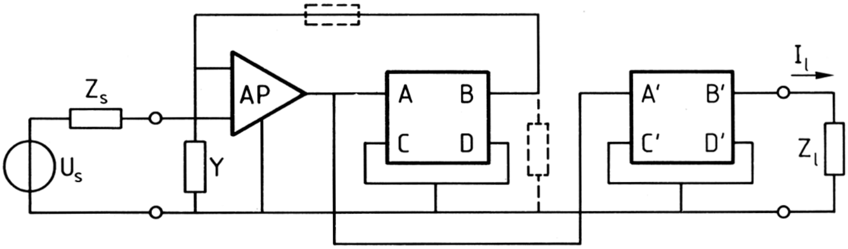


Figure 1.25 Indirect-feedback transadmittance amplifier.

(iv) The technique of indirect voltage sensing

A technique similar to that described above might he used to realise amplifiers with indirect voltage sensing for fixing either the transmission parameter A or C. however, the sign of the voltage-gain factor or transimpedance is the same as that of an amplifier with direct voltage sensing, unless equivalent inverting and non-inverting output stages are available. These output stages should have zero output impedances in the ideal case. This desired combination of properties cannot be easily obtained. Fortunately, there seems to be no urgent need for these amplifier types.

(v) The technique of indirect voltage comparison

To illustrate the operating principle of amplifiers with an indirect comparison of the feedback voltage and the signal-source voltage, a configuration with *output shunt feedback* is shown in figure 1.26.

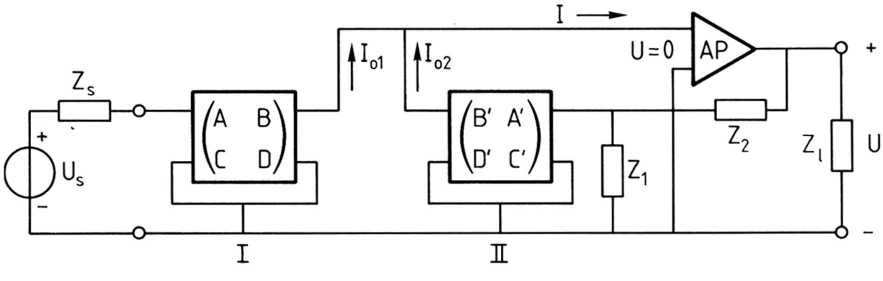


Figure 1.26 Inverting voltage amplifier with indirect voltage comparison.

Two input stages are needed. Their transfer properties are described again with the transmission parameters. Assuming the loop gain in the direct-feedback loop (formed by input stage II, the active part AP and the voltage divider Z_1-Z_2 to be infinite, the voltage U and the current I are necessarily zero. The asymptotic value of the voltage gain then becomes:

$$A_{u\infty} = \frac{U_\ell}{U_s} = -\,(1 + Z_2/Z_1)\left(\frac{B' + D'Z'}{B + DZ_s}\right) \tag{1.30}$$

where $$Z' = \frac{Z_1 Z_2}{Z_1 + Z_2}$$

Note that the parameters A and C (characterising the voltage-gain factor and the transimpedance, respectively) ideally do not play any role in the amplifier transfer. Unlike direct-feedback amplifiers, but like the technique with indirect current sensing, two parameters (B and D) of the input stages influence the transfer function. Because these parameters are signal dependent, the transfer function is non-linear even when loop gain is infinite. For obtaining satisfactory linearity compensation techniques may be necessary.

The ratio of the parameters B and D represents the input impedance of the input stages with short-circuited output. Non-linearities can therefore alternatively be ascribed to the signal-induced input impedance variations.

A necessary condition for obtaining perfect linearity by compensation is that $Z' = Z_s$ The currents i_{o1} and i_{o2} have equal magnitudes, but opposite signs. As a consequence, either *linear, symmetrical or complementary* transfer characteristics of the input stages are additionally required. The parameters B and B' as well as D and D' then can have equal voltage dependencies. when both conditions are met, the voltages u_o and u_s have the same forms, but opposite signs, and a linear and accurate voltage amplifier results.

An obvious disadvantage of meeting the condition $Z' = Z_s$ is that both input stages equally contribute to the total amplifier noise. The amplifier-noise contribution is therefore 3 dB higher than it would be when determined by stage I exclusively. A second disadvantage is that the source impedance must be well known. The noise contribution of stage II can be reduced, by reducing its transadmittance by connecting several devices in series.

A general configuration with the equivalent input noise sources of stages I and II is given in figure 1.27. The parameter D of stage II does not change significantly by the series connection, but the parameter B is enlarged by a factor n, provided that the bias conditions in all stages are equal.

The voltage gain of the configuration of figure 1.27 is given by:

$$A_{u\infty} = \frac{U_\ell}{U_s} = -n(1 + Z_2/Z_1)$$

provided that the influence of the parameter D can be disregarded. When the power-density spectrum of the input noise voltage u_{nI} of stage I is denoted as $S(u_{n\mathrm{I}})$, stage II will have an input noise voltage characterised by a spectrum about n times as large: $S(u_{n\mathrm{II}}) \approx nS(u_{n\mathrm{I}})$. The input noise current sources $i_{n\mathrm{I}}$ and $i_{n\mathrm{II}}$ have equal spectra. The spectra of the output noise currents of the stages due to the input noise voltages are given by:

$$S(i_{o\mathrm{I}}) = \frac{S(u_{n\mathrm{I}})}{B^2} \quad S(i_{o\mathrm{II}}) = \frac{nS(u_{n\mathrm{I}})}{n^2B^2} = \frac{S(u_{n\mathrm{I}})}{nB^2}$$

The relative contribution to the total equivalent input noise voltage is therefore smaller as n increases. The influence of the noise current source $i_{n\mathrm{II}}$ can be kept small by taking a small value for Z'.

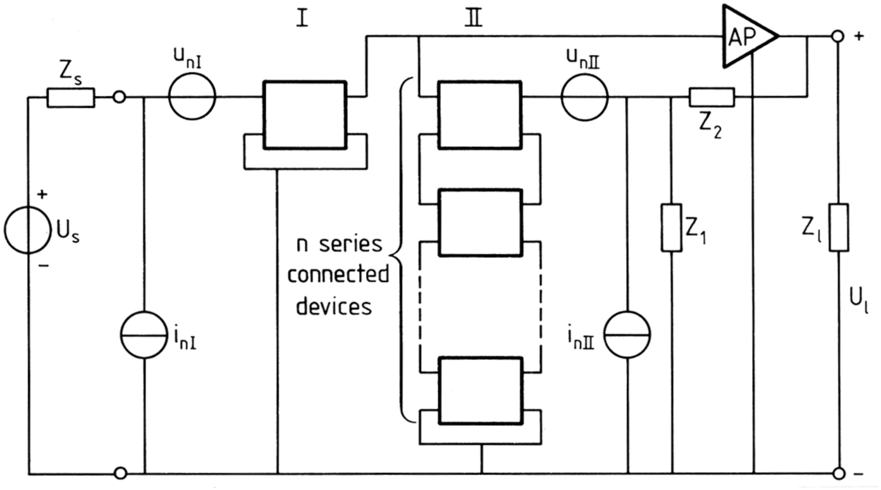


Figure 1.27 Indirect-feedback voltage amplifier with equivalent input noise sources.

We can conclude that, when the gain scaling is achieved by series connection of configurations, noise performance need not significantly be deteriorated. The limits of the amplifier noise performance can thus be almost completely set by the noise properties of stage II (provided that the noise contribution of the active part is small).

(vi) Input-stage implementation

Two simple input-stage implementations are shown in figure 1.28. Complementary configurations are used in figure 1.28a, whereas the differential pairs of figure 1.28b have symmetrical transfer functions. The latter configuration can be made to have an inverting as well as a non-inverting input.

With small values of the source impedance and of Z', the terms with D and D' in (1.14) may be sufficiently small to be disregarded. The remaining non linearity may arise from unequal variations in the stage transconductances $1/B$. It should be noted that unequal bias currents in the devices generally result in non-linearity. For maintaining linearity it would be necessary for the transadmittance of the stages to be linearly proportional to the driving input voltages. In bipolar transistors this is not the case, but when field effect devices are used in the input stages, linearity is hardly harmed by unequal currents because the above condition is roughly met. FET input stages are favourable also because the parameter D is much smaller, at least at low frequencies. Linearity can be improved alternatively by applying, for example, negative feedback to the input stages. Further considerations with respect to input-stage implementation will be put off until later chapters.

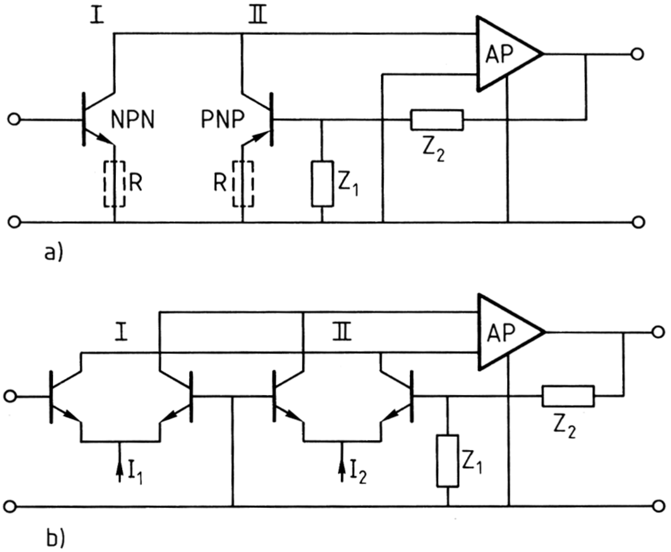


Figure 1.28 Simple implementations of input stages in indirect-feedback voltage amplifiers. Configuration a) has complementary input stages. Configuration b) has symmetrical input stages.

(vii) Indirect voltage comparison combined with output series feedback

Instead of realising low output impedances by using output shunt feedback, it is equally possible to obtain high output impedances by means of output series feedback. A basic configuration of such a transadmittance amplifier is shown in figure 1.29. The active part AP must have a floating output port, the same as in conventional transadmittance amplifiers. Assuming an ideal active part, the transadmittance of the amplifier becomes:

$$Y_\infty = \frac{I_\ell}{U_s} = Y\left(\frac{B + D/Y}{B + DZ_s}\right) \qquad (1.31)$$

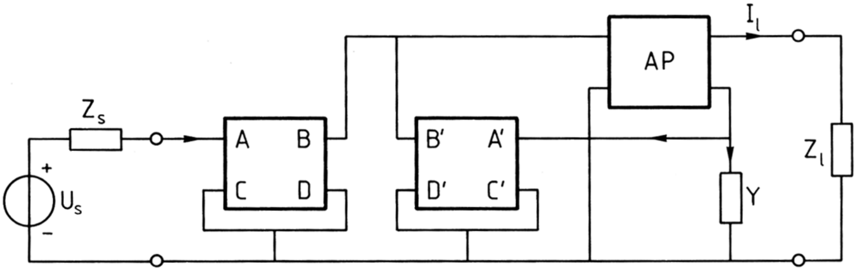


Figure 1.29 Transadmittance amplifier with indirect voltage comparison.

The parenthesised factor has the same form as in expression (1.30), and the discussions with respect to that expression apply equally well to this configuration.

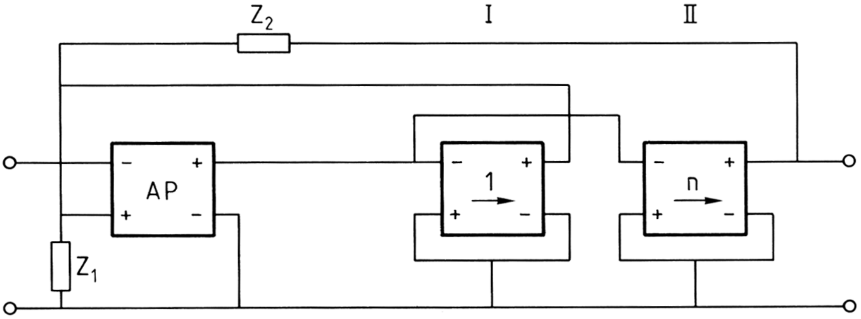
(viii) The technique of indirect current comparison

A technique similar to that described above might be used to realise amplifiers with indirect current comparison, fixing either the parameter *C* or *D*. However, the signs of the transimpedance or current-gain factor thus obtained are equal to those of the corresponding direct-feedback configurations, unless equivalent inverting and non-inverting input stages with very low input impedances are available. These properties are difficult to obtain but there seems to be no immediate need for these configurations.

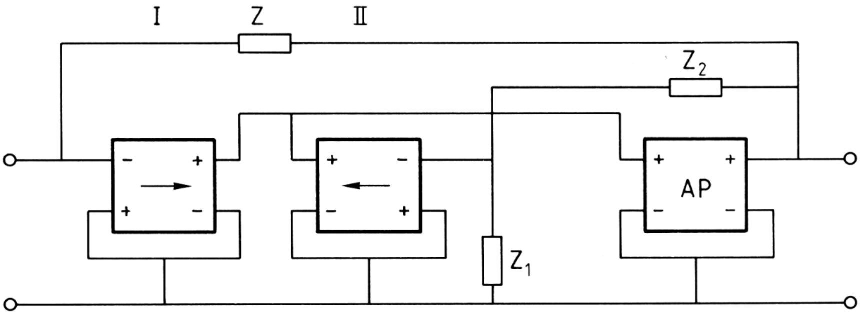
(ix) Dual-loop indirect-feedback configurations

In the foregoing we have shown that it is possible to realise inverting accurate and linear current- and voltage-gain factors by means of indirect feedback as well as non-inverting transadmittances. By fixing a second transfer parameter of such indirect-feedback configurations by means of a direct-feedback loop, it becomes possible to obtain an accurate input or output impedance at one port in combination with a zero or infinite impedance at the other port. Table 1.3 in section 1.4.2 shows the various combinations of transmission parameters that have to be fixed for this purpose. Figure 1.30 shows basic configurations of amplifier types 6, 7, 10 and 11.

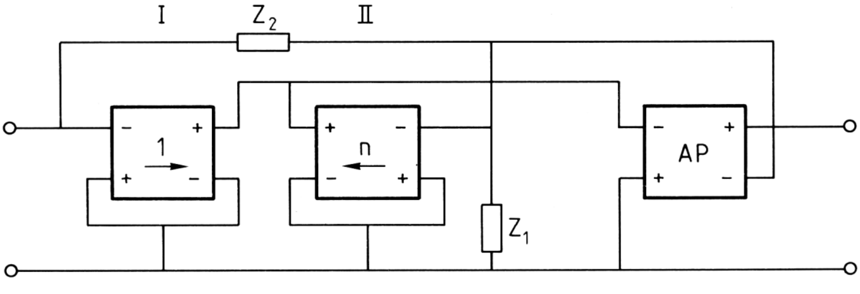
The arrows indicate the directions of the signal transfer. The examples are believed to be self evident.



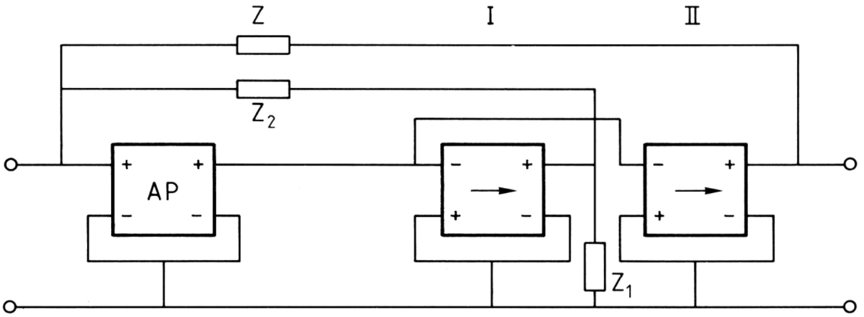
(a) realisation of configuration 6 $B = Z_1/n$ $A = 1/(1 + Z_2/Z_1)$



(b) realisation of configuration 7 $A = -1/(1 + z_2/Z_1)$ $C = -1/Z$



(c) realisation of configuration 10 $B = Z_1/n$ $D = 1/(1 + Z_2/Z_1)$



(d) realisation of configuration 11 $C = -1/2$ $D = -1/(1 + Z_2/Z_1)$

Figure 1.30 Indirect-feedback configuration with an additional direct-feedback loop for the realisation of dual-loop amplifier types.

In the foregoing it has been demonstrated that indirect-feedback techniques offer useful possibilities for realising amplifier configurations that cannot be obtained with conventional (no transformers) passive feedback networks. The active-feedback techniques to be discussed in the next section offer similar possibilities. They may yield better approximations, however, of the ideal situation, where the transmittance is determined by passive elements exclusively.

1.5.3 Active-feedback amplifiers

The use of an active-feedback technique may be considered when a transmittance with a sign opposite that of the corresponding passive-feedback configuration is needed. It can serve as an alternative for an indirect-feedback configuration. This technique also offers the possibility to realise the dual-loop amplifiers 6, 7, 10 and 11 of table 1.3, section 1.4.2. We will not go into much detail in dealing with the possible configurations, but will only give some examples.

Figure 1.31 illustrates the principle of operation of an active-feedback configuration. The transmission parameter D has been fixed in this case by using an inverting current attenuator as a feedback element. In order to realise an amplifier with a linear and accurate transfer function, the inverting current attenuator itself should be a negative-feedback circuit, either with direct feedback or with indirect feedback.

In section 1.4.4 we have seen that it is not possible to realise an approximately non-energic negative-feedback inverting current amplifier (or attenuator) while using a passive feedback network. Making an inventory of possible configurations of the feedback network, we find:

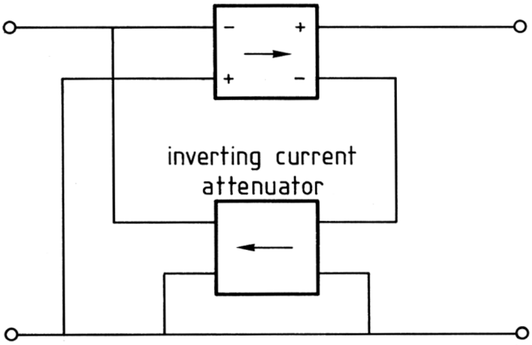


Figure 1.31 Inverting current amplifier with active feedback.

- A transimpedance amplifier with a series impedance at its output;
- A transadmittance amplifier with a shunt impedance at its input;
- An indirect-feedback current amplifier (with a current-gain factor smaller than unity).

The first two configurations are shown in figure 1.32. Assuming nullor properties of the active parts in both the forward path and in the feedback network, the current-gain factors of both amplifiers have values $\alpha = -Z_2/Z_1$.

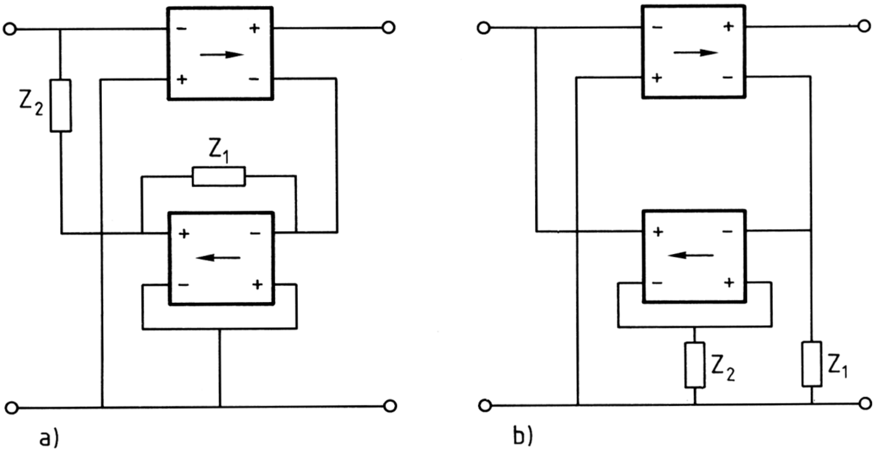


Figure 1.32 Two inverting current amplifiers with active feedback. $\alpha = -Z_2/Z_1$.

Though the active feedback networks basically have a non-linear transfer function, their influence on amplifier non-linearity may be rather small when the current and voltage variations at the output of the feedback network remain small. The output voltage is kept small by taking a small Z_1. For optimum noise performance, however, the impedances Z_2 should have values as large as possible (as will be elucidated in chapter 3). A compromise between noise and distortion performance may therefore be necessary.

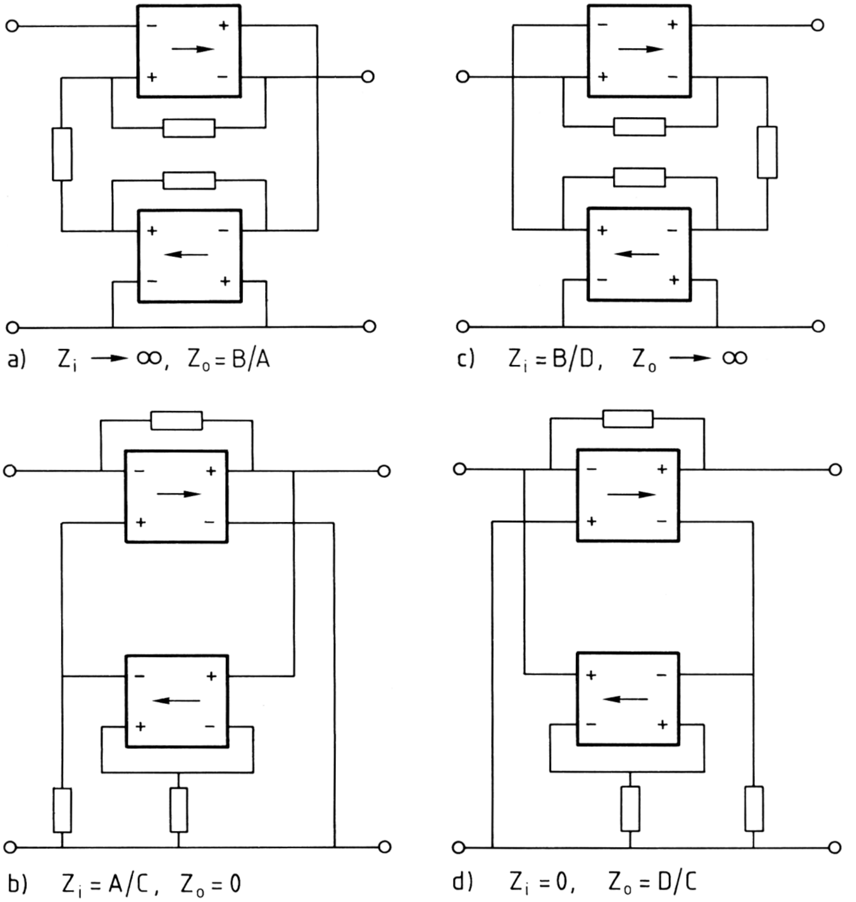
Raising the loop gain in the active part of the active feedback network may, of course, improve its linearity. Stability considerations will finally be decisive for its allowable complexity.

Similar considerations can be given for other single-loop active-feedback configurations. An important criterion for their usefulness is the extent to which they behave like approximations of non-energic feedback amplifiers. In other words, the question is can they perform better than 'brute-force'-terminated amplifiers, where series and/or shunt impedances at the input or output are used? Suffice it to note here that they can, under certain conditions. We will address this question in some more detail in chapters 3 and 4.

The active-feedback technique enables us to realise inverting voltage and current amplifiers and non-inverting transimpedance and transadmittance amplifiers. In addition, there is a possibility to realise amplifiers with an accurate impedance at one port and an (ideally) infinite or zero impedance at the other port. Figure 1.33 shows some basic configurations, which can be found with the aid of table 1.3, section 1.4.2. The arrows within the active-part symbols indicate the directions of the signal transfer. The active feedback networks in figure 1.33 are realised with the basic configurations of figure 1.21. Alternatively, indirect-feedback attenuators might be used.

1.6 Discussion

In the first section of this chapter we formulated some criteria for the optimum adaptation of amplifiers to signal sources and loads. For this purpose we made use of the concept 'best reproducing relation' and 'transmittance', to emphasise that the required amplifier properties for a given source and load follow from information-transfer aspects. In order to realise accurate and linear transfer of information, the application of negative feedback around an active amplifier part is imperative. For the impedance adaptation to sources and loads, either very low, very high or accurate and linear input and output impedances are necessary. Nine different types of transmittances which can match all types of sources and loads can be distinguished.



a) $Z_i \rightarrow \infty$, $Z_o = B/A$

c) $Z_i = B/D$, $Z_o \rightarrow \infty$

b) $Z_i = A/C$, $Z_o = 0$

d) $Z_i = 0$, $Z_o = D/C$

Figure 1.33 Basic amplifier configurations with active feedback loops in addition to passive loops.

A first introductory, largely theoretical approach to the realisation problem made use of ideal feedback two-ports: the ideal transformer and the ideal gyrator. The ideality of these two-ports refers to the absence of series and shunt impedances in the amplifier signal path, which would deteriorate signal-to-noise ratio and efficiency. ‘Brute force’ terminations are pernicious, and therefore, the proper types of feedback at input and output are needed.

It was shown that 16 basically different configurations exist, one having no feedback loops (the nullor), four with a single loop, six with two loops, four with three loops and one with four loops. Each feedback loop determines one transmission or transfer parameter nearly independently, the reason why a description of the amplifier transfer properties in terms of these parameters is preferred. All desired transmittances can be realised with these basic configurations. The configurations with three feedback loops have no obvious applications.

More practical configurations are found when transformers and impedances are admitted as elements of the feedback networks. This class of feedback configurations, presented in this chapter, is believed to be new. Some of the configurations have been incidentally proposed before, but a systematic classification such as given here has not been found elsewhere. Most configurations with passive impedances, except transformers, in the feedback network are well known, though they have not been classified systematically before either.

Some transmittances cannot be obtained with this class of amplifiers because one of the transfer parameters must have the opposite sign. The application of active feedback offers the possibility to realise the desired additional amplifier types. Two fundamentally different active-feedback methods can be distinguished. The first has been called an indirect-feedback technique because both sensing and comparing are not related to the actual load and source quantities. Accurate device matching is required for this technique, and therefore it is suitable for applications in integrated circuits.

The second is called an active-feedback technique. Sensing and comparing are direct in this case, where one of the conventional passive-feedback configurations or an indirect-feedback configuration is employed in the feedback network.

The basic amplifier configurations presented in this first chapter provide the designer with the possibility to select the right amplifier type for his specific design problem.

Unfortunately, the feedback network generally degrades noise performance as well as power efficiency to some extent. In the subsequent chapters we will address the question of how far these quality aspects are deteriorated. Furthermore, we will deal in detail with the design of the active part of the amplifier.

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2

A Classification and some Properties of Configurations with one Active Device; some Active-feedback Stages

2.1 Introduction

In chapter 1 we introduced the asymptotic-gain model, describing the transfer function of a negative-feedback amplifier in terms of its asymptotic gain and its loop gain. The asymptotic gain is found by attributing nullor properties to the active part of the amplifier. It can be regarded as the design goal. With the aid of the finite loop gain $A\beta$, the departures from the ideal transfer are described. Once the asymptotic gain is fixed by the selection of the proper basic amplifier configuration and by the determination of the feedback network, the design of the active part of the feedback loop can be started.

The design of a high-performance amplifier requires a careful selection of the correct types of stages in the active part of the amplifier. A justifiable choice requires knowledge of the performance of all possible stage configurations. The configuration that shows the best performance with respect to all quality aspects involved should be selected.

In this chapter we will present a classification of amplifier stages with one active device, with the aim of facilitating the selection of stage configurations in the active part of the amplifier. This classification is based on a view that the common-emitter (CE) stage (bipolar transistor) and the common-source (CS) stage (field-effect transistor) are basic *non-feedback* configurations, whereas *all other* configurations are feedback versions of the CE and CS stages. In addition we will present some configurations with an active feedback network where the feedback loop embraces one active device.

We will adopt a rather formal view in describing the transfer properties of the single-device stages. Each configuration, being considered as a two-port, will be described with four parameters.

As a matter of fact, the *environment* of a configuration determines whether all parameters have a significant influence on the transfer function or not. In former days, vacuum tubes (triode, tetrode and penthode) were usually described with two parameters by those who were engaged in low-frequency amplifier design. A description with a transconductance and a voltage gain factor was perfectly adequate because shunt impedances in the signal path (in the form of ‘load impedances’) were always present, making the current-gain factor and the transimpedance irrelevant parameters in this low-frequency region. A current-driven triode stage would have been an exceptional phenomenon. Nowadays, ‘active loads’ are widely used and they can have very high impedances. Even a common-source stage can have a relatively low input impedance at all except very low frequencies.

Analogous to the simplified description of the vacuum tube, the bipolar transistor in a common-emitter configuration is frequently characterised by two parameters as well, viz. its current-gain factor and its transconductance. A further example is the current mirror, where usually a great deal of attention is paid primarily to the current-gain *factor,* whereas the influence of the other transfer parameters on the *actual* current transfer is frequently ignored. Such simplified descriptions of active devices with less than four parameters are believed to give at the very least an incomplete picture of design possibilities and at the very worst a distorted insight into design problems.

We will describe the transfer properties of amplifier stages with the aid of transmission or transfer parameters because such a description will appear to be extremely useful for finding design criteria for optimum noise performance, accuracy, linearity and high-frequency behaviour. Besides, it corresponds very well to the concept that all configurations except the CE and CS stages are feedback configurations. The transfer parameters will in turn be expressed in the parameters of equivalent circuits in order to facilitate associations with physical effects.

The classification is given for bipolar transistors only. Similar results are found for corresponding FET configurations.

2.2 Equivalent circuits of active devices

Before focusing on the above-mentioned classification, we will first discuss the equivalent circuits of the active devices which will be used in further design considerations. These equivalent circuits should be as simple as possible but still reasonably allow for various physical effects. A description with matrix parameters exclusively does not meet this requirement.

2.2.1 Equivalent circuit of the bipolar transistor

The hybrid-π model is considered to be particularly suitable for describing the small-signal behaviour of a bipolar transistor. It is depicted in figure 2.1. Apart from the feedback, introduced in CE operation by the capacitance $C_{b'c}$, and by a very large resistance which is ignored, the model is unilateral. Moreover, the model comprises only one controlled source. In addition to the parameters of this model, some other parameters will be used to characterise the device properties, related to those of the hybrid-π model in the following way:

$$\alpha_{eo} = g_e r_{b'e}, \quad \mu_{eo} = g_e r_{ce}, \quad g_e = \frac{1}{r_e} = \frac{qI_C}{kT}, \quad \omega_T = \frac{1}{r_e(C_{b'e} + C_{b'c})}$$

where ω_T is the current-gain-factor transit frequency. The less usual parameter μ_{eo} has values between 10^3 and 10^4.

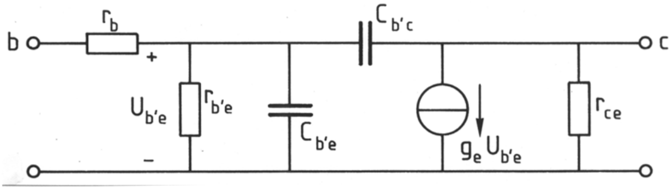


Figure 2.1 Hybrid-π model of the bipolar transistor.

2.2.2 Equivalent circuit of the field-effect transistor

The structure of the equivalent circuit of the FET is the same as that of the bipolar transistor. It is depicted in figure 2.2. Here also, some additional parameters will be used, namely:

$$\mu_{so} = g_m r_{ds}, \quad \omega_T = \frac{g_m}{C_{gs} + C_{gd}} = \frac{g_m}{C_{iss}},$$

where ω_T is the current-gain-factor transit frequency. The parameter μ_{so} has values between 50 and 500.

The transfer parameters of the single-device configurations and other transfer functions will be expressed in the parameters as given in this section.

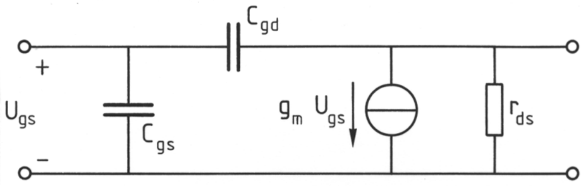


Figure 2.2 Equivalent circuit of the field-effect transistor.

2.3 Configurations with one active device; local-feedback configurations

2.3.1 Introduction

In chapter 1 it was shown that it is possible to realise 16 types of negative-feedback amplifiers with fundamentally different transfer properties using up to four feedback loops. Each feedback loop essentially fixes one of the four transfer parameters at a smaller value, while the others are completely or virtually (in the configurations with more than two loops there is some interaction) unaffected. In each of the four types of single-loop amplifiers, for example, one of the transfer parameters is reduced and fixed, while the others remain equal or very nearly equal to those of the active part, so that in practical amplifiers these remain large, non-linear and inaccurate. Of the possible amplifier stages with one active device, the common-emitter stage (bipolar transistor) and the common-source stage (field-effect transistor) have transfer parameters that are all large, non-linear and inaccurate. It should be noted that, as mentioned before in chapter 1, it is not necessary to make feedback effects in amplifiers always explicit. Though it is possible, for example, to consider the capacitance $C_{b'c}$ in a bipolar transistor as a feedback impedance, fixing the transimpedance of a CE stage at high frequencies, we prefer not to do so because this feedback effect does not produce an accurate and linear transfer parameter.

Considering, on the above grounds the CE stage and CS stage as non-feedback stages and as approximations of a nullor with two interconnected terminals, we can identify the CE stage with the nullor symbol. We can then find the other configurations from the classifications of chapter 1.

2.3.2 Classification of local-feedback stages

In figure 2.3, the identification of the CE configuration with the nullor symbol is given. To provide an initial impression of the magnitudes of the transfer parameters of the CE stage, their low-frequency values (expressed in the parameters of the hybrid-π equivalent circuit) are presented below:

$$\mu_{lf} = \frac{-\mu_{eo}}{1 + r_b/\alpha_{eo} r_e} \approx -\mu_{eo} \qquad \gamma_{lf} = \frac{-g_e}{1 + r_b/\alpha_{eo} r_e} \approx -g_e$$

$$\zeta_{lf} = -\alpha_{eo} r_{ce} = -\alpha_{eo}\mu_{eo} r_e \qquad \alpha_{lf} = -\alpha_{eo}$$

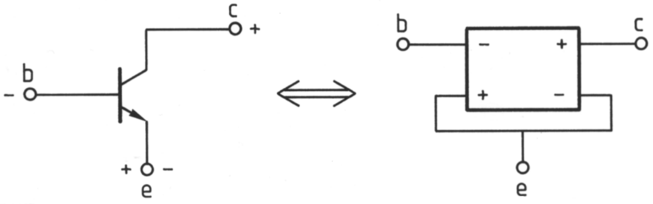


Figure 2.3 Identification of transistor symbol with the symbol of an active part.

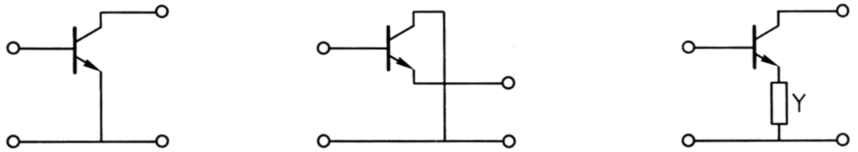
Though high-quality information transfer is, as can be seen from these parameter values, obviously not possible with a CE stage, it can in order to provide power gain be used without objections as an amplifying stage in the active part of a negative-feedback amplifier. Other types of amplifier stages, however, may be equally useful for application in the active part. In order to develop criteria regarding the utility of the various single-device configurations we will now give a classification of stages conforming to the above view, where the CE stage of figure 2.3 forms the active part of the various negative-feedback configurations. For practical reasons, we will restrict ourselves to circuits with an active device together with impedance-feedback networks, having one terminal in common with the source and load.

Because two terminals of this active part are in common, three out of the ten configurations given in chapter 1, figure 1.21 cannot be realised. All resulting single-device negative-feedback configurations with fundamentally different transfer properties are shown in figure 2.4. We will refer to them henceforth as local-feedback stages. Their more or less familiar names are given in the figure captions. Table 2.1 lists the types of feedback used in each configuration. Note that feedback-stabilised voltage-gain factors and current-gain factors cannot exceed unity in the single-loop configurations as a consequence of the restraints we have put on the circuits.

Table 2.1 Types of feedback at the input and output in single-device configurations.

	configuration	type of input feedback	type of output feedback
a.	(CE stage)	—	—
b.	(emitter follower)	series	shunt
c.	(series stage)	series	series
d.	(shunt stage)	shunt	shunt
e.	(CB stage)	shunt	series
f.		shunt, series	shunt, series
g.		shunt, series	series, shunt
h.		shunt, series	series, shunt

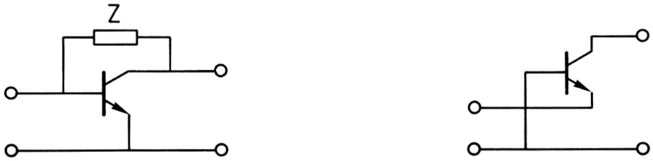
Somewhat aside, it may be noted that it is possible to influence the transfer parameters by inserting series and shunt impedances at the input and/or output of the single-device configurations. Such measures were rejected in chapter 1. Evidence in support of this rejection will be given in later chapters. Larger values of the transfer parameters can, as a matter of course, be obtained by applying positive feedback to the active part of figure 2.3. An inverting feedback network is required for this purpose. Therefore, either a transformer or an active feedback network is needed. Configurations with transformers are believed to be unpractical in wide-band amplifiers; we will exclude them from further considerations. Active positive-feedback networks can be useful in some situations, in particular for base-current compensation [1], [2]. We will deal with neither of these any further.



a) CE stage

b) CC stage
"emitter follower"

c) "Series stage"



d) "Shunt stage"

e) CB stage
"current follower"

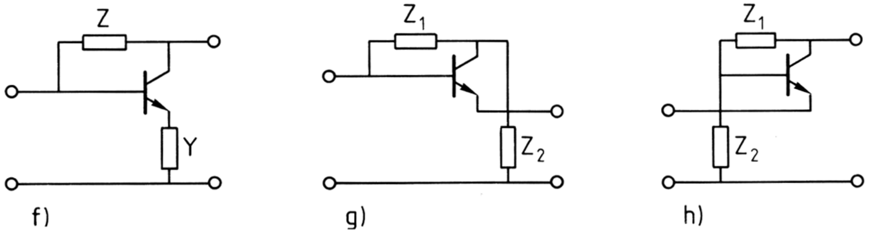


Figure 2.4 Configurations with a single active device.

2.3.3 Calculation of the transfer parameters of the local-feedback stages

The transfer parameters of the local-feedback configurations can be found either by straightforward calculation or with the aid of the asymptotic gain model. The use of this model for the calculation of a certain transfer parameter is meaningful only when the feedback factor β has a non-zero value. If the loop gain $A\beta$ is large, the parameter in question can be approximated by its asymptotic value $A_{t\infty}$.

As an example, the parameter $\alpha = (I_o/I_i)_{U_o=0}$ of a CB stage is determined. Figure 2.5 shows the equivalent circuit. In order to find the asymptotic value of the parameter α, the voltage $U_{b'e}$ is supposed to be zero, and consequently, I_b is zero. It is seen immediately that:

$$\alpha_\infty = (I_o/I_i)_{A\beta\to\infty} = 1$$

The loop gain is most easily determined from the equivalent circuit drawn in figure 2.6, where it is given by

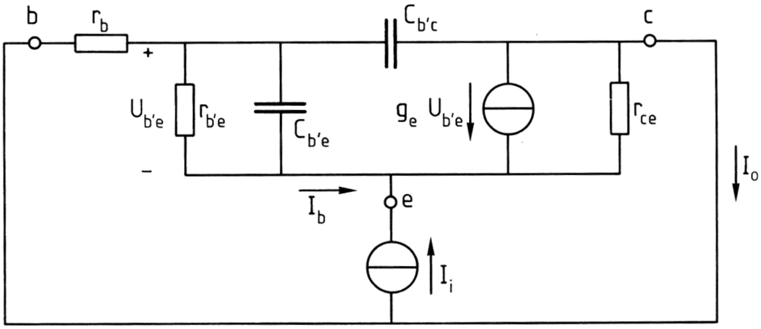


Figure 2.5 Equivalent circuit for the determination of the parameter α of a CB stage.

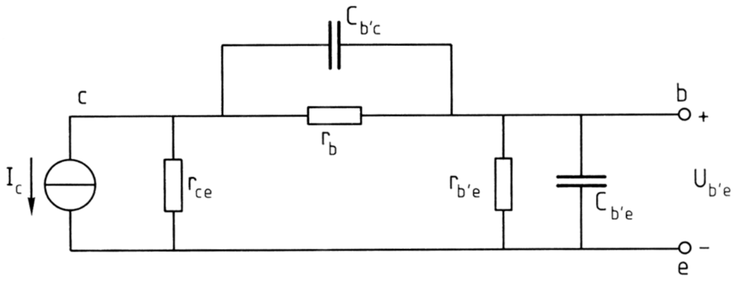


Figure 2.6 Equivalent circuit for the determination of the loop gain of a CB stage.

$$A\beta = g_e \frac{U_{b'e}}{I_c}$$

In this circuit, r_{ce} is large compared with the impedance in parallel with it, and to a good approximation we can write:

$$\beta = \frac{U_{b'e}}{I_c} = \frac{r_{b'e}}{1 + j\omega\dfrac{\alpha_{eo}}{\omega_T}}$$

so that

$$A\beta = \frac{-\alpha_{eo}}{1 + j\omega\dfrac{\alpha_{eo}}{\omega_T}}$$

With this expression for $A\beta$, the parameter α can be written as:

$$\alpha = \left(\frac{I_o}{I_i}\right)_{U_o=0} = \frac{\alpha_{eo}}{1 + \alpha_{eo}} \cdot \frac{1}{1 + j\dfrac{\omega}{\omega_T}} = \alpha_b$$

It should be emphasised that the feedback model cannot be used for the determination of the other transfer parameters of the CB stage, because when voltage driven or with open terminals, the loop gain becomes zero or is determined by the internal feedback via $C_{b'c}$. As a consequence of zero loop gain, the transfer parameters μ, γ and ζ are practically equal, apart from their signs, to the corresponding parameters of the CE stage.

2.3.4 Transfer properties of the local-feedback stages

The other configurations of figure 2.4 with a single feedback loop behave in a way similar to the CB stage. One parameter is strongly affected by the feedback action; the others are nearly equal to those of the CE configuration. In the series and shunt stages some effect on all parameters is noticeable as a result of the series and shunt impedances. The character of the non-stabilised transfer parameters is preserved, however, as far as linearity and accuracy are concerned.

The feedback configurations can be used in the active part in two different ways. First, the single-loop configurations can be driven and/or loaded so that their transfer functions become approximately equal to that of a CE stage. As building blocks of the active part of a negative-feedback amplifier, they can be useful alternatives for a CE stage, either because different input or output impedances can be realised, or because the transfer function may have the opposite sign, as is the case in the CC and CB stages. In the second place, a configuration can be used, of course, for the specific character of the feedback-stabilised parameter. It must then be driven and loaded accordingly.

In the configurations with two types of feedback at the input and output more transfer parameters are influenced by the feedback action. Table 2.2 gives a survey of the approximate low-frequency values of the transfer parameters of the given configurations. The asymptotic values are given (indicated by an asterisk) for the parameters that are basically affected by negative feedback. In the non-energic-feedback configurations (the CC stage and the CB stage) the feedback-stabilised parameters have actual values close to the asymptotic values, because the loop gain amounts to about α_e In the other configurations the loop gain will be smaller. It may be noted that several types of transmittances given in table 1.1 can be realised with these configurations. The loop gain, however, is small, and we will not pay much attention to this aspect of the local-feedback stages. Our primary aim is to establish criteria regarding the suitability of stages for application in overall negative-feedback amplifiers.

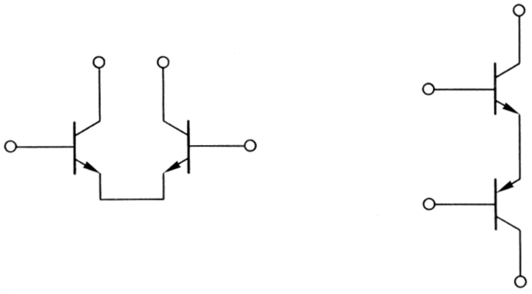
2.4 Balanced configurations

All single-device configurations as presented and discussed in the previous section can be transformed into balanced versions by combining two of them. The balanced versions of the CE stage, for example, are the differential amplifier stage and its complementary counterpart as shown in figure 1.7a and b. The use of such balanced configurations may be preferable or even imperative in certain applications. Especially in cases where signals with very low-frequency components have to be amplified (‘dc amplifiers’) differential configurations like the one of figure 1.7a are more suitable than unbalanced single-device stages because in the latter case the temperature-dependent transistor characteristics are responsible for unwanted signal components that cannot be distinguished from the wanted signal components. These temperature-dependent ‘signals’ can be largely rejected in the balanced stages. Other important applications of differential stages are those where the signal source carries a large interfering (common-mode) signal with respect to ground and where a load has to be driven, for example, from a balanced current source as in double-balanced mixers. The complementary balanced stages are frequently employed in class AB and class B output stages. The transfer parameters of the balanced stages have values very similar to those of unbalanced stages. Even order distortion in the transfer functions of these stages may be effectively suppressed so that their linearity may be improved. The noise properties of the balances stages are generally somewhat worse. Because the design considerations, with respect to the quality aspects treated in this work, of amplifiers with balanced stages are very similar to those with unbalanced stages, we will not deal with these explicitly in the following. It should be noted that there are some other important quality aspects in relation to such amplifiers (for example, common-mode rejection, offset quantities and their drift with temperature, etc.)

which will not be mentioned hereafter. The reader interested in these design aspects is referred to reference [3].

Table 2.2 Approximate low-frequency values of the transfer parameters of single device configurations. Asymptotic values are indicated by an asterisk.

	configuration	μ	γ	ζ	α
a.	(CE stage)	$-\mu_{eo}$	$-g_e$	$-\alpha_{eo}\mu_{eo}r_e$	$-\alpha_{eo}$
b.	(CC stage)	1^*	g_e	$\alpha_{eo}\mu_{eo}r_e$	α_{eo}
c.	(series stage)	$-\mu_{eo}\frac{Y_f r_{b'e}}{1 + Y_f r_{b'e}}$	$-Y^*$	$-\alpha_{eo}\mu_{eo}r_e$	$-\alpha_{eo}\frac{Y_f r_{ce}}{1 + Y_f r_{ce}}$
d.	(shunt stage)	$-\mu_{eo}\frac{Z_f}{Z_f + r_{ce}}$	$-g_e$	$-Z^*$	$-\alpha_{eo}\frac{Z_f}{r_{b'e} + Z_f}$
e.	(CB stage)	μ_{eo}	g_e	$\alpha_{eo}\mu_{eo}r_e$	1^*
f.		$1 - YZ^*$	$(1 - YZ)/Z^*$	$(1 - YZ)/Y^*$	$1 - YZ^*$
g.		1^*	g_e	$Z_1 + Z_2{}^*$	$(Z_1 + Z_2)/Z_2{}^*$
h.		$(Z_1 + Z_2)/Z_2{}^*$	g_e	$Z_1 + Z_2{}^*$	1^*



(a) differential amplifier stage (b) complementary version

Figure 2.7 Balanced versions of the CE stage.

2.5 Frequency dependency of the transfer parameters of the single-loop single-active-device configurations

In this section we will present the expressions — including their frequency dependency — for the transfer parameters of the single-active-device configurations with one feedback loop. These expressions can be used for considerations regarding noise and distortion performance to be treated in chapters 3 and 4, respectively, and as a base for bandwidth and stability considerations which will be dealt with in chapter 5.

To obtain expressions in easily surveyable form, the transistor parameters are given in admittance notation in the following way:

$$g_b = \frac{1}{r_b}, \quad Y_f = pC_{b'c}, \quad Y_i = g_{b'e} + pC_{b'e}$$

The parameter r_{ce} is not taken into account, because it complicates the expressions, and furthermore, its influence is noticeable in the low-frequency region only. The expressions can also be used for FET's, in which case $g_b = \infty$, $Y_f = pC_{gd}$ and $Y_i = pC_{gs}$, while r_{ds} is ignored.

The expressions for the transfer parameters of the CE stage are given below. The factor $(Y_f - g)$ is present in each parameter. This means that independently of the source and load conditions the transfer function of a CE stage has a zero in the right half-plane, given by:

$$n = \frac{1}{r_e C_{b'e}}$$

$$\mu = \left(\frac{U_\ell}{U_s}\right)_{I_\ell = 0} = \frac{g_b(Y_f - g)}{Y_f(Y_i + g + g_b)} = \frac{1}{A} \tag{2.1}$$

$$\gamma = \left(\frac{I_\ell}{U_s}\right)_{U_\ell = 0} = \frac{g_b(Y_f - g)}{Y_i + g_b + Y_f} = \frac{1}{B} \tag{2.2}$$

$$\zeta = \left(\frac{U_\ell}{I_s}\right)_{I_\ell = 0} = \frac{Y_f - g}{Y_f(Y_i + g_b)} = \frac{1}{C} \tag{2.3}$$

$$\alpha = \left(\frac{I_\ell}{I_s}\right)_{U_\ell = 0} = \frac{Y_f - g}{Y_i + Y_f} = \frac{1}{C} \tag{2.4}$$

Though this zero has a slightly unfavourable influence on stability in negative feedback amplifiers because it causes some excess phase shift, its influence will generally be negligible because it is much larger than f_T. Because r_{ce} was neglected, μ and ζ have poles in the origin. The actual values of these poles can, however, easily be estimated by bearing in mind that the low frequency values of μ and ζ have approximate values $-\mu_{eo}$ and $-\alpha_{eo} r_{ce}$, respectively, while for low frequencies μ and ζ, as given by (2.1) and (2.3), can be approximated by $\mu = -1/[pC_{b'c}(r_e + r_b)]$ and $\zeta = -1/pC_{b'c}$. The frequency for which the magnitudes of the corresponding parameter values are equal determines the actual pole locations.

For several purposes an approximation of the transfer parameters is permitted where the right-half-plane zero is neglected and only the low-frequency pole (in the case of μ and ζ) is taken into account. Furthermore, Y_f will generally be small with respect to $Y_i (C_{b'c} \ll C_{b'e})$. The transfer parameters can then be written as:

$$\mu = \frac{-g g_b}{(g + g_b) pC_{b'c}}, \qquad \gamma = \frac{-g g_b}{g_{b'e} + g_b + pC_{b'e}}$$

$$\alpha = \frac{-g}{g_{b'e} + pC_{b'e}}, \qquad \zeta = \frac{-1}{pC_{b'c}}$$

The expressions for the transfer parameters of the single-loop local-feedback configurations are given in table 2.3, while table 2.4 presents the approximate expressions where the transfer zeros and the non-dominant poles are neglected.

The *CC stage* appears to have a common factor in the numerator of two transfer parameters given by $(Y_i + g)$. This means that unless the load impedance becomes infinite, a zero is found in the left half-plane, approximately at $n = -\omega_T$. In negative-feedback amplifiers this zero can have a favourable influence on stability because it can cause some phase-lead compensation.

The *CB stage* generally appears to have complex zeros in the left half-plane independently of source and load conditions according to the common factor $[g(g_b + Y_f) + Y_f Y_i]$ in the numerator of all transfer parameters. The real parts of these zeros are equal to $\omega_T/2$, while the imaginary parts are dependent on ω_T, $C_{b'c}$ and r_b. It must be noted that these zeros are found in the case where the feedback capacitance is active between the nodes b′ and c in the hybrid-π equivalent circuit, whereas they are absent in the case where this capacitance works between the nodes b and c. According to some authors (e.g. [4]) the feedback capacitance in the equivalent circuit must be split up into two parts, one taking hold of the external base contact b and the other of the internal base b′. Anyhow — their imaginary parts being generally much larger than their real parts — these zeros — if present at all — do not have considerable influence in whatever situation, except when a resistance is taken intentionally in series with the base. Then the imaginary part becomes smaller and the zeros can have a beneficial influence on the stability.

Table 2.3. High-frequency transfer parameters of single-device feedback stages.

configuration	μ	γ	ζ	α
CE	$\frac{g_b(Y_f - g)}{Y_f(Y_i + g + g_b)}$	$\frac{g_b(Y_f - g)}{Y_i + g_b + Y_f}$	$\frac{Y_f - g}{Y_f(Y_i + g)}$	$\frac{Y_f - g}{Y_i + Y_f}$
CC	$\frac{g_b}{Y_f + g_b}$	$\frac{g_b(Y_i + g)}{Y_i + Y_f + g_b}$	$\frac{1}{Y_f}$	$\frac{Y_i + g}{Y_i + Y_f}$
series	$\frac{g_b(Y'_f - g)}{Y_f(Y_i + g'_b + g)}$	$\frac{g_b(Y' - g)}{Y_i + g'_b + Y'_f}$	$\frac{Y'_f - g}{Y_f(Y_i + g)}$	$\frac{Y'_f - g}{Y_i + Y'_f}$
shunt	$\frac{-gg_b + (Y_i + g_b)Y'_t}{(Y_i + g_b)Y'_t}$	$\frac{gg_b + (Y_i + g_b)Y'_t}{(Y_i + g_b)}$	$\frac{-gg_b + (Y_i + g_b)Y'_t}{Y'_t g_b(Y_i + g)}$	$\frac{-gg_b + (Y_i + g_b)Y'_t}{(Y_i + g_b)Y'_t + Y_f g_b}$
CB	$\frac{g(g_b + Y_f) + Y_f Y_i}{Y_f(Y_i + g + g_b)}$	$\frac{g(g_b + Y_f) + Y_f Y_i}{Y_i + g_b + Y_f}$	$\frac{g(g_b + Y_f) + Y_f Y_i}{g_b Y_f(Y_i + g)}$	$\frac{g(g_b + Y_f) + Y_f Y_i}{(g + Y_i)(g_b + Y_f)}$

Table 2.4 *Approximate expressions for the transfer parameters of single-device configurations. Feedback-stabilised parameters are indicated by an asterisk.*

configuration	μ	γ	ζ	α
CE	$-\frac{gg_b}{g+g_b}\cdot\frac{1}{Y_f}$	$-\frac{gg_b}{g_b+Y_i}$	$-\frac{1}{Y_f}$	$-\frac{g}{Y_i}$
CC	$1\ *$	$\frac{gg_b}{g_b+Y_i}$	$\frac{1}{Y_f}$	$\frac{g}{Y_i}$
series	$-\frac{gg_b}{g+g'_b}\cdot\frac{1}{Y_f}$	$-\frac{gg_b}{Y'_b+Y_i}\ *$	$-\frac{1}{Y_f}$	$-\frac{g}{Y_i+Y'_f}$
shunt	$-\frac{gg_b}{g+g'_b}$	$-\frac{gg_b}{g_b+Y_i}$	$-\frac{1}{Y'_t}\ *$	$-\frac{g}{Y_i+Y_t}$
CB	$\frac{gg_b}{g+g_b}\cdot\frac{1}{Y_f}$	$\frac{gg_b}{g_b+Y_i}$	$\frac{1}{Y_f}$	$1\ *$

In the *series stage* with a real feedback impedance $R_f = 1/Y_f$ a zero is found in the right half-plane just as with the CE stage. However, the zero is smaller because of the reduced transconductance:

$$n = \frac{1}{C_{b'c}(r_e + R_f)}$$

Moreover, this particular way of applying feedback causes a pole and a zero in all transfer parameters in the left half-plane. They are only slightly different, yet complicate high-frequency design. They can be made coincident, however, by shunting the feedback resistor R_f with a capacitance C_f, their product being given by $R_f C_f = 1/\omega_T$. For the transfer parameters as given in table 2.4 and in all cases where the series stage will be discussed, this pole-zero cancellation is assumed to be accomplished. The expressions for the transfer parameters of the series stage in that case are the same as those of the CE stage. The parameters Y_f and g_b however, are replaced in some places by parameters Y'_f and g'_b given respectively by:

$$Y'_f = \frac{R_f + r_e}{r_e}\cdot Y_f \quad \text{and} \quad g'_b = \frac{R_f + r_e}{r_e}\cdot g_b$$

The enlarged value of g'_b is connected with the reduced and stabilised transconductance; the enlarged value of Y'_f generally brings about an inferior high frequency behaviour as will appear in chapter 5.

The transmission parameters of the *shunt stage* can be approximated to easily surveyable expressions by supposing that the internal feedback admittance Y_f shunts the external feedback admittance Y_t. The resulting admittance is indicated as Y'_t ($Y'_t = Y_f + Y_t$). This approximation can be allowed as long as Y_t is small with respect to [5]. A zero in the right half-plane is found at nearly the same place as in a CE stage if the feedback admittance is real.

2.6 Amplifier stages with local active feedback

In chapter 1 we dealt with configurations having active feedback networks and distinguished between active- and indirect-feedback techniques. By using these techniques, amplifier configurations can be obtained with transfer parameters having signs opposite to those of direct-feedback configurations. It becomes possible to realise inverting current- and voltage-gain factors as well as non-inverting transadmittances and transimpedances. Moreover, accurate and linear impedances at one amplifier port can be combined with an infinite or zero impedance at the other port.

Active feedback networks can, of course, also be applied to *one* active device. The various possibilities can be found again in a way analogous to that in chapter 1. Some indirect-feedback configurations frequently encountered in monolithic integrated circuits may be worth mentioning.

2.6.1 Amplifier stages with indirect local feedback

(i) Configurations with indirect current sensing

Among the simplest realisations of amplifier stages with indirect current sensing are the well-known current mirror and Gilbert's current-gain cell [6]. These circuits can be easily derived from the basic configurations in chapter 1. Figure 2.8 depicts both configurations. The amplifier symbol is dashed in order to elucidate the derivation.

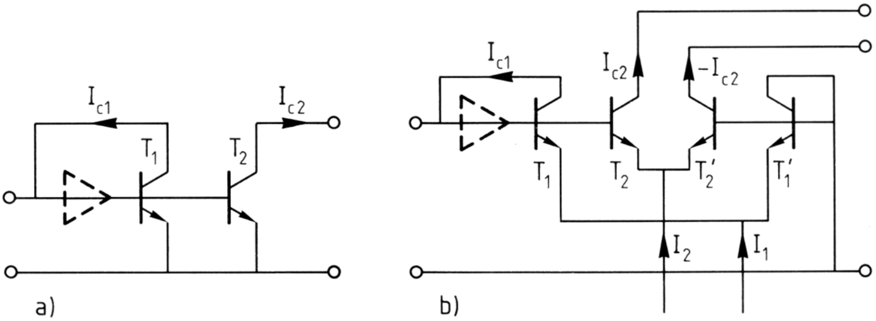


Figure 2.8 *Current mirror and a single-sided-input version of Gilbert's gain cell, derived from the general configurations in chapter 1.*

The output currents of the transistors T_1 are completely fed back to the inputs, so that they are nearly equal to the input currents. The transfer parameters α therefore nearly equal unity, when the bias currents of all transistors are equal.

In the case of bipolar transistors, the collector current depends exponentially on the base-to-emitter voltage. Consequently, the transconductance is linearly proportional to the bias current I_C. Thanks to this unique property it is possible to obtain a current gain larger than unity by merely scaling up the bias currents in transistors T_2. Accuracy and linearity of the transfer parameter n are preserved in that case. The simple scaling possibility thus relies on a particular mechanism viz. The stated exponential relationship. With other types of devices such as FET's a linear gain larger than unity is possible only when the currents as well as the number of devices are scaled to the same extent.

It should be noted that gain setting by bias-current scaling does not seem to be possible for the circuit of figure 2.8a without scaling the number of devices. This is only true when the feedback is active for both signal and bias currents. Figure 2.8a represents a signal path, however, and bias currents in T_1 and T_2 can, in principle, be chosen independently. For very low-frequency applications scaling of the number of devices (or alternatively, of the emitter area) is necessary indeed because the signal current in unbalanced configurations cannot be distinguished from the bias current. The balanced version equipped with bipolar transistors as in figure 2.8b is advantageous because scaling up the current I_2 for realising a larger gain will do.

Inverting as well as non-inverting transfer parameters can be obtained with the circuit of figure 2.8b. Moreover, the gain can be controlled. A differential input is possible in addition to the differential output [6].

It may be clarifying to compare the current-mirror performance with the corresponding configuration with direct feedback (the CB stage). For this purpose, the approximate values of the low-frequency parameters of the current mirror are given below:

$$\mu \approx -\mu_{eo} \qquad \zeta \approx -r_{ce}$$

$$\gamma \approx -\gamma_e \qquad \alpha \approx -1$$

The important differences with the CB-stage parameters are the negative signs and the much lower value of the transimpedance, indicating that the influence of the load impedance on the current gain is substantially larger. The direct current-sensing technique is obviously superior to indirect sensing.

The parameters of the configuration with a current gain larger than unity can be estimated by considering that the transconductance is proportional to the bias current, r_{ce} is roughly inversely proportional and μ is roughly independent of the bias current. Hence, for a current-gain factor n:

$$\mu \approx -\mu_{eo} \qquad \zeta \approx -r_{ce}/n$$

$$\gamma \approx -ng_e \qquad \alpha \approx -n$$

where g_e and r_{ce} are parameters of the local-feedback transistor. For the current-gain cell we find parameters with about the same values.

A simple circuit with a rather accurate and linearised output impedance, corresponding to that of figure 1.30d is depicted in figure 2.9. If $I_2 = nI_1$, the transfer parameters are approximately given by:

$$\mu \approx -g_e Z_f \qquad \zeta \approx -Z_f$$

$$\gamma \approx -ng_e \qquad \alpha \approx -n$$

The output impedance is approximately equal to Z_f/n, when the circuit is driven from a relatively high-impedance source. The output power loss in Z_f can be small, when the circuit is used for characteristic impedance matching.

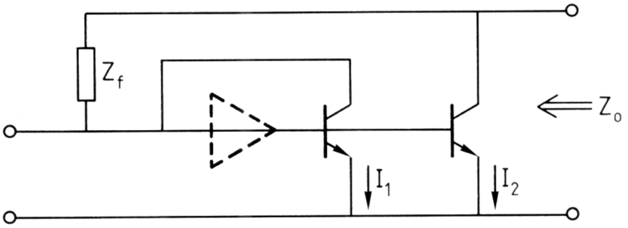


Figure 2.9 Configuration with an accurate output impedance and a low input impedance.

(ii) Configurations with indirect voltage comparison

Local-feedback configurations with indirect voltage comparison are less well known. Bilotti [7] has proposed a gain-stabilisation technique of this kind. The configuration can be derived from figure 1.28a by replacing the amplifier symbol by a short circuit. Figures 2.10 and 2.11 depict examples of indirect-feedback voltage-gain stages. In the circuit of figure 2.10a, the combination of T_2 and the resistances R_1 and R_2 is known as a U_{BE} multiplier. The circuit of figure 2.10b is a very obvious one. The transfer parameters of these stages are approximately given by:

$$\mu \approx -n \qquad \zeta \approx -\alpha_{eo} n r_e$$

$$\gamma \approx -g_e \qquad \alpha \approx -\alpha_{eo}$$

where $n = 1 + R_1/R_2$ for the configuration of figure 2.10a, while $n = 3$ (equal to the number of 'diodes') in figure 2.10b. A 'voltage mirror' with bipolar transistors, according to figure 2.10b, but with one 'diode' can handle only a small voltage of a few hundred mV's. Compared to the corresponding configuration with direct voltage comparison, the emitter follower, its transimpedance is much smaller, with the consequence that the source impedance has a much stronger influence on the voltage transfer. The balanced configuration of figure 2.11 with normally-off MOSFET's can handle much larger input signals and has a very high input impedance in spite of the lower value of the transimpedance ζ as compared with a source follower. A simple circuit with an accurate and linear input impedance is shown in figure 2.12. It is based on the stage of figure 2.10a. The transfer parameters are approximately given by:

$$\mu \approx -n \qquad \zeta \approx -Z_f$$

$$\gamma \approx -g_e \qquad \alpha \approx -\alpha_{eo}$$

with $n = 1 + R_1/R_2$. The input impedance is therefore approximately equal to Z_f/n.

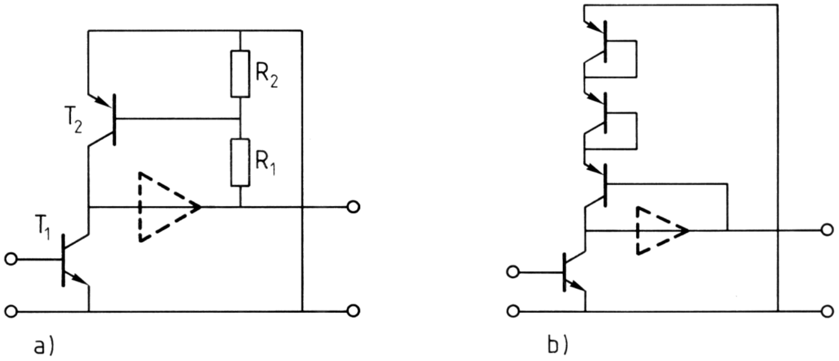


Figure 2.10 Indirect-feedback voltage-gain stages

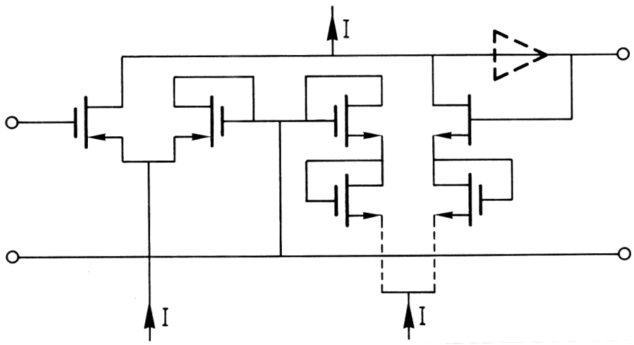


Figure 2.11 Indirect-feedback voltage-gain stage with MOSFET's.

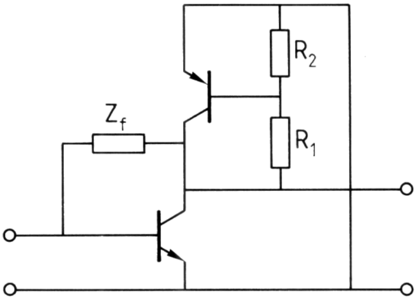


Figure 2.12 Configuration with low output impedance and accurate input impedance ($Z_i \approx Z_f/(\ell + n)$, $n = 1 + R_1/R_2$).

2.6.2 Example of a configuration using an active-feedback technique

The well-known Wilson current mirror [8] depicted in figure 2.13 is an example of a configuration using an active-feedback technique around a single device. The output current of this device is sensed directly by the input port of an indirect-feedback configuration, namely a simple current mirror. Input current and feedback current are compared directly as well. The output impedance of the Wilson current mirror is much higher than that of the simple mirror as a consequence of the direct current sensing. The configuration has transfer parameters with the same order of magnitude as a CB stage. The sign of the parameters, however, is opposite.

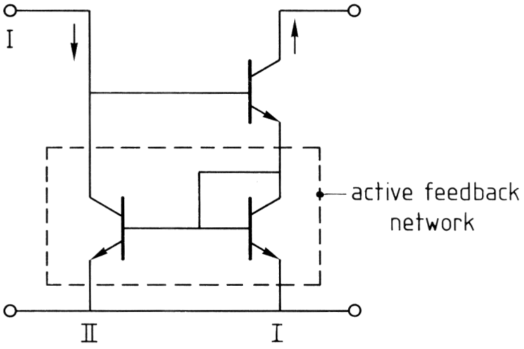


Figure 2.13 Three transistor current mirror, where the active feedback network (dashes) is equipped with an indirect-feedback current amplifier.

Instead of a unity-gain current mirror in the feedback path it is of course possible to use an indirect-feedback current attenuator in order to realise a current gain larger than unity. The configuration can then be provided with a second passive loop for fixing the output impedance at an accurate and linear value, the same as in figure 2.9.

2.7 Input- and output-stage configurations for floating active parts

Active parts with floating input ports require the use of input- or output-stage configurations with an input or output port, respectively, well isolated from its environment. The extent to which a non-perfect isolation plays a role in transfer accuracy and non-linearity will be examined in chapter 4. Here we will give some simple examples of input and output configurations that may offer a reasonable amount of isolation. Figure 2.14 shows these bipolar transistor and junction-FET input and output configurations. The input ports are indicated as (i–i′); and the output ports as 0–0′, respectively.

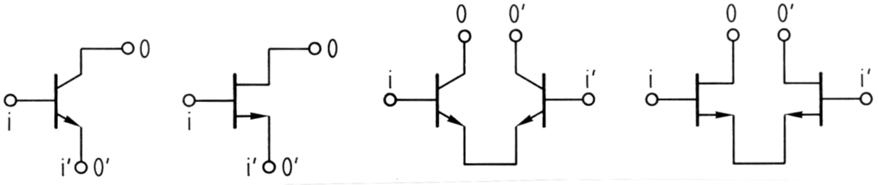


Figure 2.14 Possible input and output configurations of the active part for realising floating input or output ports.

2.8 Discussion

The classification of single-device configurations, as given in the previous sections is extremely useful for the sharp formulation of design criteria, as will appear in the subsequent chapters.

A striking feature of the characterisation of stages with transmission parameters is that all parameters of a local-feedback stage are nearly equal to those of a CE stage except the parameter(s) affected by feedback. The feedback-stabilised parameters can be approximated by the value following from the asymptotic-gain model. So — when the expressions for the CE-stage parameters are known — the values of the parameters of the local-feedback stages can be easily found approximately, though sufficiently well for various design considerations. It is noted somewhat aside that the classification given here also shows how to realise transmittances with cascaded local-feedback stages. Each stage should be driven and loaded for that purpose in such a way that it has a reasonable amount of loop gain. As a result, the transmittance of the whole amplifier will approximately be given by the product of the individual transmittances of the local-feedback stages. Though in this way amplifiers can be realised with relatively little design effort, we will not deal with their design in detail because their performance is suboptimum compared with overall negative-feedback amplifiers.

As can be seen from the loop equivalent circuit — as given, for example, in figure 2.15 for a series stage with load and source impedances — the loop gain has a maximum value of about α_{eo} in the case of a bipolar transistor and of μ_{so} in the case of an FET. This maximum is reached only when source and load impedances are either zero or infinite and when impedances for realising series and shunt feedback are large and small, respectively.

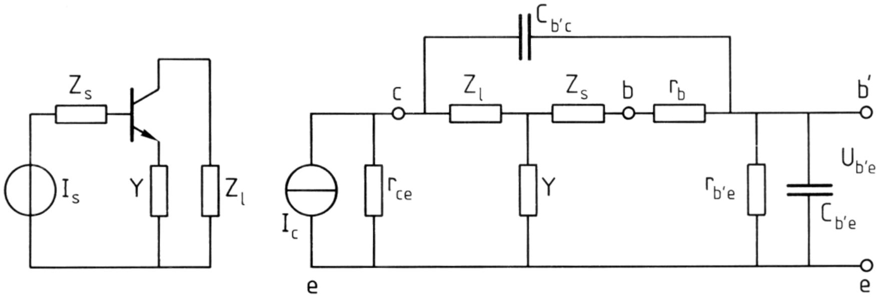


Figure 2.15 Series stage with source and load impedances together with the loop equivalent circuit ($A\beta = g_e U_{b'e}/I_c$).

Unless concessions are made, for example with respect to noise performance, it will usually be impossible to realise optimum information transfer with a cascade of local-feedback stages because the magnitudes of source or load impedances impede the realisation of a sufficiently large loop gain.

For the indirect-feedback configurations as presented in this chapter, it may be less obvious to make the negative feedback explicit. Nevertheless, we prefer such an approach, because it provides a useful basis for comparison of stages with direct and indirect feedback. Moreover, the stages can be derived from the more general configurations that were given in chapter 1.

The use of these types of stages is not necessarily restricted to integrated circuits, as may be clear from some examples given. Where possible, preference must be given to stages with direct feedback because of their superior transfer properties. Further support for this preference will be given in chapters 3 and 4, where noise and distortion performance are investigated.

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3

Design Considerations for Optimum Noise Performance of Negative-Feedback Amplifiers

3.1 Introduction

Apart from the desired signal information in the output signal of an amplifier there are undesirable signal components as well, caused by hum, interference, drift, noise, distortion, etc. A number of these unwanted signal components have their origin outside the signal path of the amplifier and can, in principle, be arbitrarily well suppressed. We will concisely deal with some of them in chapter 6.

Random noise, however, is inherent in the use of active devices and is thus generated in the signal path. Its influence can be minimised but can never be reduced to zero. The magnitudes of the noise sources depend on the type of device and its bias conditions. The magnitudes of the equivalent input noise sources of the amplifier are determined, in principle, by all noise sources in the amplifier. The best situation, however, occurs when the noise contribution is determined exclusively by the noise sources of the active device in the input stage, while at the same time this noise contribution is minimised with respect to a given source impedance.

The optimisation of the noise performance of overall negative-feedback amplifiers will in this chapter be approached purely from a designer's point of view. The only purpose of the procedure to be developed here is to minimise the *noise contribution* of amplifiers.

A description of amplifier noise performance in terms of one or another figure of merit, such as noise figure, noise measure, noise temperature, signal-to-noise ratio, etc. is neither necessary nor very appropriate in this context. Nevertheless, following common practice, noise figure (F) and signal-to-noise ratio will be used now and then to characterise the noise performance of an amplifier with a given signal source.

In order to design an amplifier with optimum noise performance together with an accurate and linear information transfer, various design aspects have to be considered, namely:

- The influence of series or shunt impedances on noise performance.
- The influence of feedback networks on noise performance.

- The noise production of various input-stage configurations.
- The noise contribution of the second amplifier stage.

Before dealing with these aspects we will introduce in section 3.2 some transformation techniques as design tools that will appear to be extremely useful for fast noise analysis and hence for design.

In the third section of this chapter we will show that impedances in series or in parallel with the signal-source impedance generally deteriorate noise performance in wide-band amplifiers, even when they are lossless. Consequently, such impedance-adaptation techniques — characterised as brute-force techniques in chapter 1 — should not be used for the adaptation of the input impedance of the amplifier to the signal source.

Subsequently, in section 3.4 we will investigate the influence of feedback networks on noise performance, making distinction between non-energic, lossless and lossy feedback networks. Simple rules will ensue for dealing with this influence for various single-loop basic feedback configurations.

The classification of single-device stages given in the previous chapter, will serve as a guide for finding the optimum input configuration. Noise performance of the local-feedback stages will be compared with the CE-stage performance.

The influence of the second-stage noise will be investigated by transforming the second-stage noise sources to the input, using a transmission-parameter representation of the input stage. It will be shown that this influence can be minimised by using a non-feedback stage (CE stage, or CS stage) at the input.

The second part of this chapter presents a design procedure based on the conclusions of the former sections. It aims at an optimisation of the noise match of a properly selected amplifier type to various types of signal sources. This procedure includes correctly selecting the type of active component in the input stage and its bias conditions.

3.2 Design tools

Some useful transformations that greatly simplify noise analysis and consequently facilitate design will be treated first.

The first transformation technique deals with voltage sources and is known as the Blakesley transformation or e-shift [1]. It is illustrated in figure 3.1.

A voltage source u_n can be shifted across a node into two other branches connected to that node without affecting the mesh equations. Both networks in figure 3.1 are therefore equivalent at their terminals. As a matter of course, the sources u_n are fully correlated in the case where u_n is a noise source.

A similar transformation applies to current sources according to figure 3.2. The current source i_n can be split into two equal sources without affecting the node equations. Both networks in figure 3.2 are therefore equivalent at their terminals.

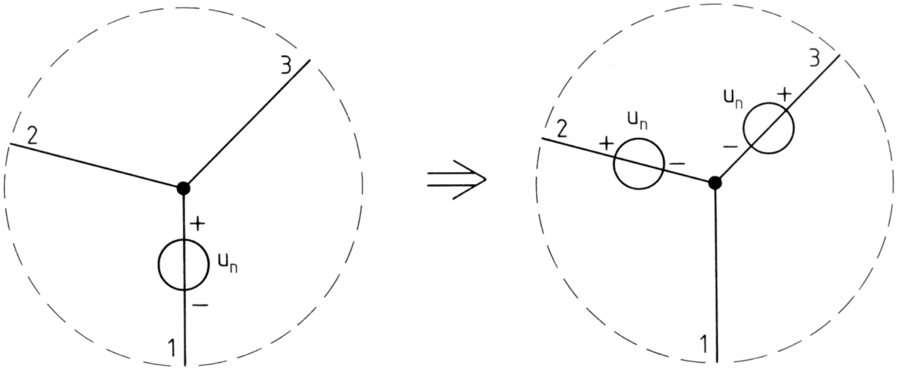


Figure 3.1. Blakesley transformation or e-shift.

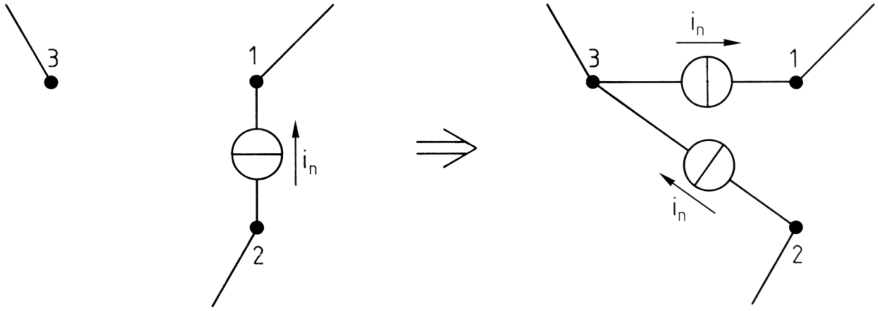


Figure 3.2 Splitting of current sources.

With the aid of the above transformations together with the well-known Norton-Thévenin conversions, noise sources somewhere within the equivalent circuit of a two-port can be transformed to sources in series or in parallel with the input or output ports.

In order to find the total equivalent noise source in series or in parallel with the input-signal-representing voltage or current source, the output sources of a two-port have to be transformed back to the input. For this purpose, a transmission-parameter representation of the two-port is most appropriate [2].

The equations of a two-port with two (noise) sources at its output, as depicted in figure 3.3, can be written as:

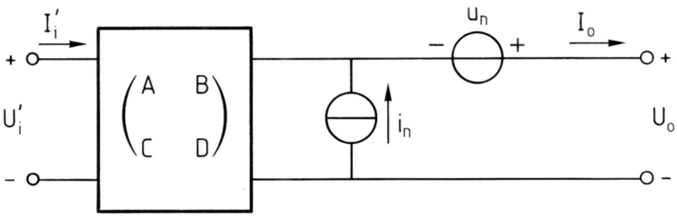


Figure 3.3 Two-port represented by its transmission parameters with noise sources at its output.

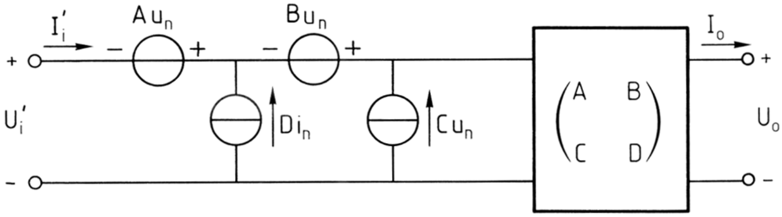


Figure 3.4 Equivalent circuit with the noise sources of figure 3.3 transformed to the output.

$$U'_i = AU_o + BI_o - AU_n - Bi_n$$

$$I'_n = CU_o + DI_o - Cu_n - Di_n$$

By rewriting these equations we easily find the equivalent *input* noise sources of the two-port:

$$U'_i = AU_n + BI_n = AU_o + Bi_o$$

$$I'_n = CU_n + DI_n = Cu_o + Di_o$$

The circuit of figure 3.4 is therefore equivalent to that of figure 3.3.

3.3 The influence of shunt and series impedances on noise performance

Initial design considerations, given in chapter 1, regarding the proper selection of a basic amplifier configuration for a given source and load were based, among other things, on the assumption that the insertion of series or shunt impedances at the input of an amplifier — in order to realise the required input impedance — may deteriorate the signal-to-noise ratio (SNR). This method for fixing driving-point impedances was therefore designated as a brute-force method. In this section we will analyse the influence of series and shunt impedances on noise performance in order to support the design considerations of chapter 1.

For the analysis we can distinguish four relevant situations, indicated in figure 3.5. The signal source is represented either by a current or a voltage source. In order to determine the SNR, with either series or shunt impedances at the amplifier input, the equivalent noise sources u_n and i_n of the amplifier have to be transformed into current or voltage sources in parallel or in series with the signal source, respectively. These transformations can be performed with the tools given in section 3.2.

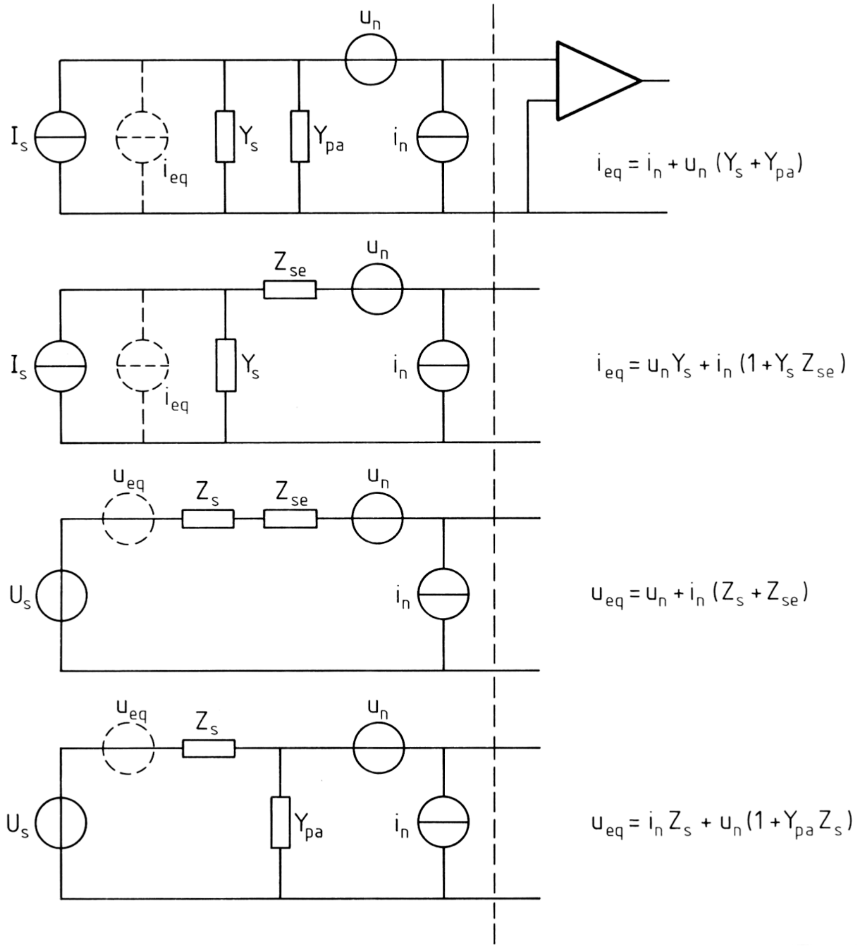

u_n
I_s
i_{eq}
Y_s
Y_{pa}
i_n
$i_{eq} = i_n + u_n (Y_s + Y_{pa})$
Z_{se}
u_n
I_s
i_{eq}
Y_s
i_n
$i_{eq} = u_n Y_s + i_n (1 + Y_s Z_{se})$
u_{eq}
Z_s
Z_{se}
u_n
U_s
i_n
$u_{eq} = u_n + i_n (Z_s + Z_{se})$
u_{eq}
Z_s
u_n
U_s
Y_{pa}
i_n
$u_{eq} = i_n Z_s + u_n (1 + Y_{pa} Z_s)$

Figure 3.5 Current- and voltage-source representation with shunt impedances Y_{pa} and series impedances Z_{se}; u_n and i_n are the equivalent input noise sources of the amplifier.

An important remark to make here is that the input impedance of the amplifier need not be known in order to calculate the SNR. The transformations are performed for the circuits to the left of the dashes. *Obviously, the transfer of signal power has nothing to do with noise performance.*

The mere presence of the shunt and series impedances leads to additional terms in the expressions for i_{eq} and u_{eq}, as given in figure 3.5. The thermal noise produced in these impedances when they are not lossless is not taken into account. These thermal noise contributions can be counted in the sources i_n or u_n, thus increasing their values and deteriorating the SNR.

From the expressions for i_{eq} and u_{eq} it can be seen that only in the case where resonance circuits are formed may the influence of one of the sources be reduced by the presence of the series or shunt impedances. Such a reduction can be put to advantage in narrow-band amplifiers and sometimes in wide-band amplifiers but only in a part of the frequency band.

When the series impedance or shunt admittance is used for impedance adaptation with the information transfer from source to amplifier in mind, their values must be large over the whole frequency range compared to those of the source. Noise performance will always be deteriorated in that case, because the influence of one of the equivalent noise sources will be increased. This is true in the case of *lossless impedances* as well as in that of series or shunt *resistors*. The latter produce in addition their own thermal noise.

The use of series and shunt impedances for the purpose of optimising the signal transfer must be rejected on the above grounds. The alternative — the use of the proper type of negative feedback at the input — will be investigated in the next section.

3.4 The influence of feedback networks on noise performance

3.4.1 Introduction

We will now investigate the influence of feedback networks on noise performance by comparing the equivalent input noise sources of the negative feedback amplifier with those of its active part which for this purpose will be assumed to be uncorrelated. We will deal with correlated sources in section 3.7. These equivalent input sources will be approximately equal to those of the first stage provided that this stage is properly designed. First-stage design considerations will be given in later sections.

The active parts of the amplifiers will be assumed to behave as nullors in this section. If they don't, the expressions for the equivalent input noise sources of the amplifier become far more complicated even though their values are not significantly different in practical situations. Therefore, the nullor approach is believed to be a justifiable simplification, which helps to improve insight.

First, we will examine amplifiers with passive feedback networks, and then we will pay some attention to active feedback networks.

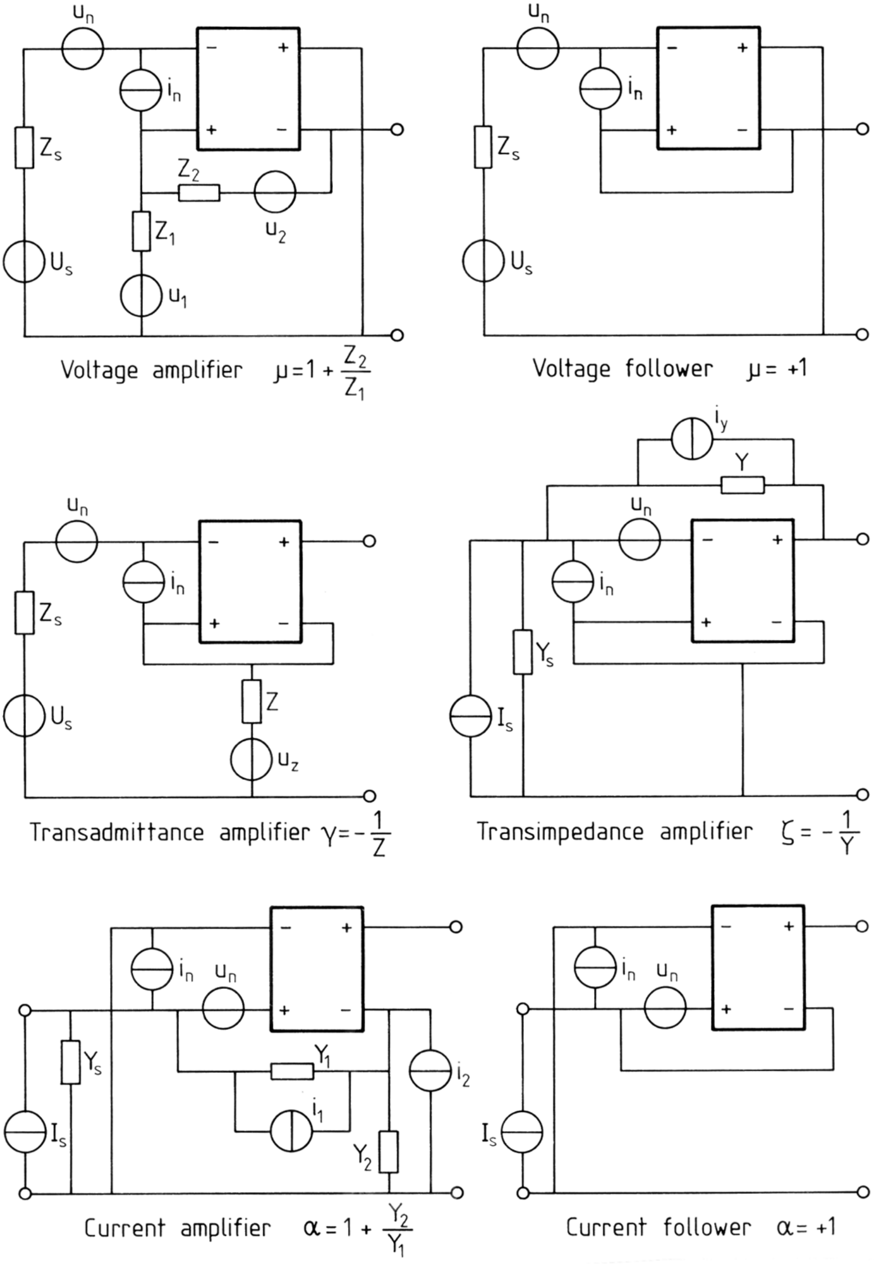
3.4.2 The influence of passive feedback networks on noise performance

The concept of *non-energic-feedback* has been introduced in chapter 1 for the classification of amplifier types with fundamentally different transfer properties. The non-energic elements — the ideal transformer and gyrator — were used as feedback networks. We have shown in chapter 1 that the equivalent input noise sources of a non-energic-feedback voltage amplifier are *exactly equal* to those of the active part. This is also true for the other single-loop nonenergic-feedback amplifiers. In the non-energic dual-loop amplifiers one of the equivalent noise sources of the amplifier is even somewhat smaller than the corresponding one of the active part.

Practical single-loop configurations are usually realised with the aid of lossless and/or lossy feedback networks. The equivalent input noise sources of such configurations are not equal to those of the active part, as will be demonstrated in the following. Figure 3.6 shows the six single-loop amplifiers, four of which have impedances in their feedback networks, and two of which are followers. These are essentially non-energic feedback configurations.

Each amplifier type is given in connection with the type of signal source to which it is best suited. The equivalent input noise sources of the active parts and the noise sources associated with the feedback impedances are given in the figures. The transformations required for finding the equivalent input noise source in parallel with the signal source are given in figure 3.7 as an example for a current amplifier. The signs of the sources are preserved only as far as necessary for taking correlations into account.

The source i_n already has its desired position. The source u_n has to be transformed together with the sources i_1 and i_2 associated with the feedback impedances. In figure 3.7a the sources i_1 and i_2 are converted into voltage sources. In figure 3.7b the e-shift has been performed on the source u_n. A further Blakesley transformation is performed in figure 3.7c on the sources u_n and u_1. The sources u'_n and u'_i (equal to u_n and u_i, respectively) can now be disregarded because of the infinite values of μ and ζ of the active part; $u_n + u_1 + u_2$ can be transformed into one current source i. In figure 3.7d this current source is split. The source i' at the output port of the active part can be disregarded because γ and α of the active part have infinite values. The source i'' can be transformed into two input sources (figure 3.7e). The resulting current source has a value i''/α_∞, where α_∞ is the current-gain factor of the amplifier ($\alpha_\infty = 1 + Y_2/Y_1$). The resulting voltage source equals zero because of the infinite value of γ.



Voltage amplifier $\mu = 1 + \frac{Z_2}{Z_1}$

Voltage follower $\mu = +1$

Transadmittance amplifier $\gamma = -\frac{1}{Z}$

Transimpedance amplifier $\zeta = -\frac{1}{Y}$

Current amplifier $\alpha = 1 + \frac{Y_2}{Y_1}$

Current follower $\alpha = +1$

Figure 3.6 *Single-loop negative-feedback amplifiers. u_n and i_n are the equivalent input sources of the active part. The noise of the feedback impedances is indicated as well.*

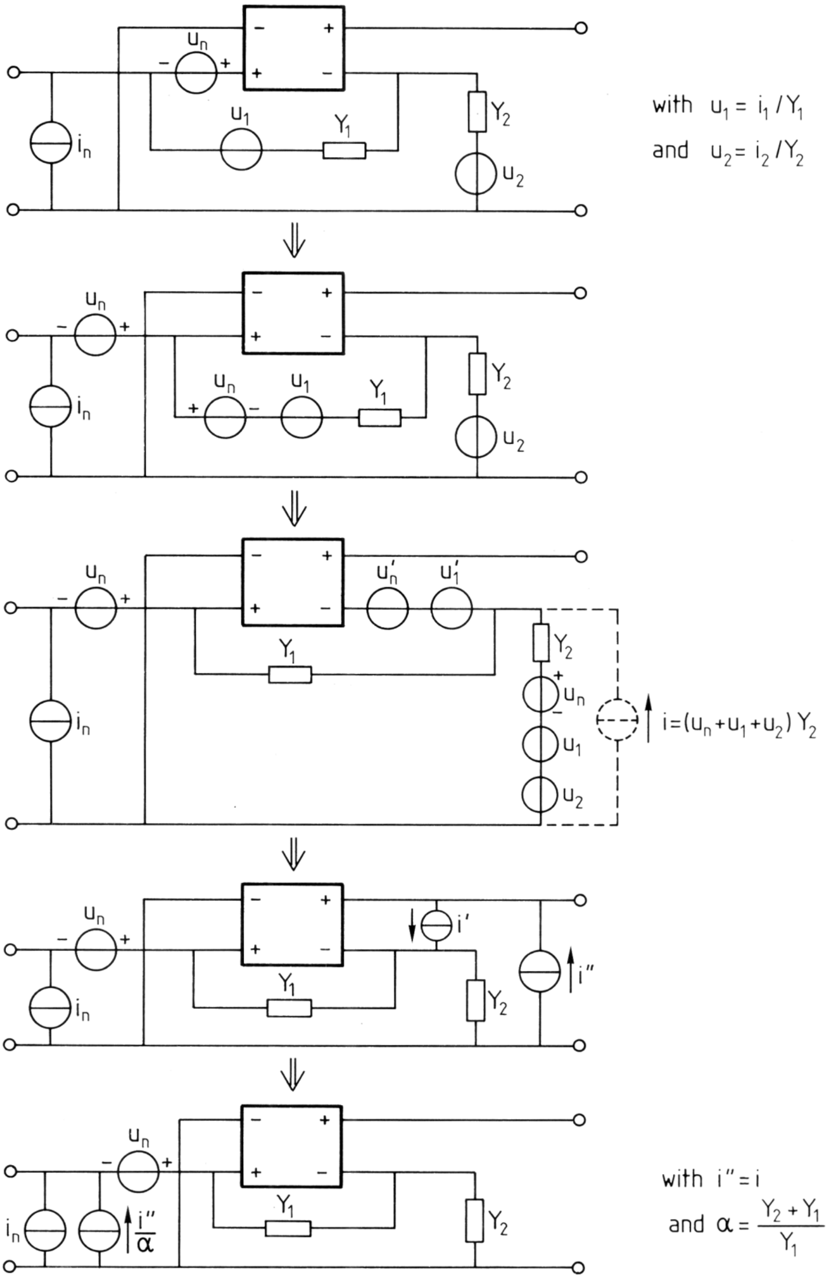


Figure 3.7 Transformations of the noise sources of the active part of a current amplifier into input noise sources of the amplifier.

A final transformation leads to one equivalent noise source in parallel with the signal source. The result is shown in figure 3.8.

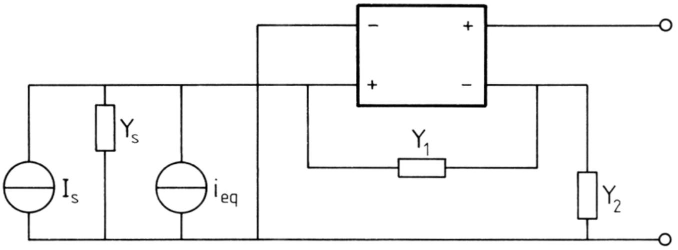


Figure 3.8 Result of the transformation of all noise sources to the input.

When the spectral densities of the noise sources are denoted as $S(u_n)$, $S(i_n)$, $S(i_1) = 4kTRe(Y_1)$ and $S(i_2) = 4kTRe(Y_2)$, the spectrum of the equivalent source i_{eq} is given by:

$$S(i_{eq}) = S(i_n) + S(u_n)\,|Y_s + Y'|^2 = 4kTRe(Y_s + Y') \tag{3.1}$$

where $Y' = Y_1Y_2/(Y_1 + Y_2)$.

From this expression it can be concluded that the feedback admittances Y_1 and Y_2 change the noise performance of the amplifier with respect to that of the active part *just as if their series connection were shunting the input terminals* and the source admittance.

The feedback network of a current amplifier thus deteriorates the noise performance in the same way as (discussed in the previous section) a shunt impedance Y_p at the input would. In the latter case we would find an equivalent input source with the same expression for its spectrum, with the exception of Y' which has to be replaced by Y_p.

We can conclude that when a low input impedance is required for high-quality information transfer, the input-shunt-feedback solution is much more favourable than the brute-force termination with a large Y_p because Y' can be much smaller than Y_p to obtain the same input impedance, and consequently, noise performance is deteriorated much less.

Similar transformations performed on the other single-loop feedback configurations of figure 3.6 lead to the following conclusions:

- The influence of the impedances Z_1 and Z_2 in the *voltage amplifier* can be accounted for as if their parallel connection were in series with the input of the active part.
- In the *voltage follower* the input noise sources are equal to those of the active part.
- The influence of the impedance Z in the *transadmittance amplifier* can be calculated as if it were in series with the signal source.
- The admittance Y in the *transimpedance amplifier* influences noise perfor-

mance as if it were in parallel with the source admittance.

- The equivalent input noise sources of the *current follower* are equal to those of the active part.

These results are exact in the case of infinite loop gain when the active part behaves as a nullor. They are good approximations in practical cases. The transformation techniques discussed in section II can be used to arrive at exact results if so desired.

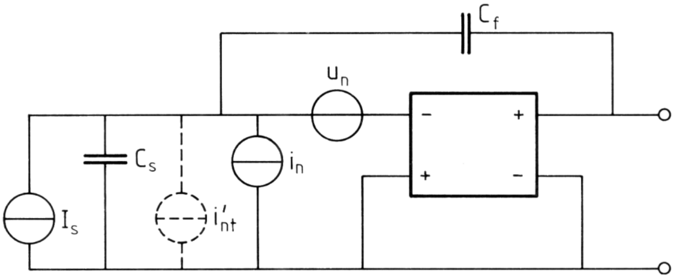


Figure 3.9 Noise sources in an integrating transimpedance amplifier (or charge amplifier) with a capacitive source.

We will conclude this section by giving an example of SNR deterioration in the case of a lossless feedback network. Figure 3.9 depicts an integrating transimpedance amplifier (or charge amplifier) in connection with a capacitive transducer. The input noise sources are assumed here to be uncorrelated. We will deal with correlated sources in later sections.

The total noise current in the case where C_f were not present amounts to

$$i_{nt} = i_n + u_n j\omega C_s$$

The associated power-density spectrum $S(i_{nt})$ of i_{nt} can be written as:

$$S(i_{nt}) = S(i_n) + S(u_n)\omega^2 C_s^2$$

With C_f present, the spectrum of the total equivalent noise current i'_{nt} has a larger value due to the enhanced influence of the source u_n. According to the above rules, we find:

$$i'_{nt} = i_n + u_n j\omega(C_s + C_f)$$

$$S(i'_{nt}) = S(i_n) + S(u_n)\omega^2(C_s + C_f)^2$$

The SNR has decreased through the presence of C_f by just the same amount as if C_f had been in parallel with the input of the active part. In the latter case, however, the influence of the shunt capacitance on the input impedance would have been much smaller.

If, with a similar amplifier, the source impedance would have been inductive, resonance effects might have improved the SNR in a small frequency band. As mentioned before, in wide-band amplifiers this can seldom be put to advantage.

3.4.3 The influence of active feedback networks on noise performance

The influence on noise performance of active feedback networks is more complicated to analyse than that of passive feedback networks because the active part of the network will generally contribute to the equivalent input noise sources. One of the design objectives is to keep this noise contribution as small as possible.

As an example we will deal first with an active-feedback current amplifier, using a transadmittance configuration together with an input shunt impedance as its feedback network. Though there is no need to use resistors in the network, not even to obtain a frequency-independent response, we will use them here for the sake of simplicity. Figure 3.10 shows the circuit with the noise sources of both active parts and of the resistors. Both active parts are assumed to behave as nullors as far as their transfer properties are concerned.

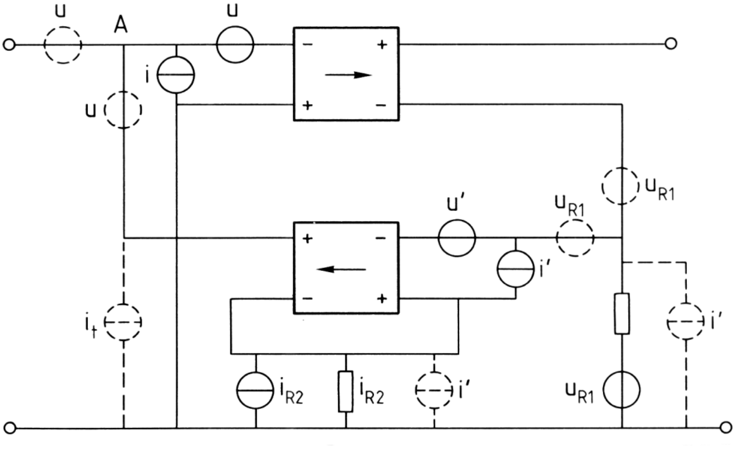


Figure 3.10 Active-feedback current amplifier with transadmittance feedback network and noise sources. The dashed sources result from transformations which are believed to be self-evident.

Because the active feedback network has an infinite output impedance, the source u can be shifted across node A without causing an additional current source. The output noise current source i_t of the active feedback network consists of the current i_{R_2}, i' and $(u' + u_{R_1})/R_2$, where it is assumed that $R_1 \ll R_2$.

In the corresponding passive-feedback configuration only R_2 would significantly contribute to the input noise current. In the active-feedback case the equivalent input noise current of the active part makes an additional contribution, which may be significant in the case of a bipolar-transistor input stage, while the source $(u' + u_{R_1})/R_2$ will usually he negligible. In section 3.5 we will study the noise properties of single-device stages in some detail. The reader is referred to that section for more detailed design criteria for optimum noise performance of the active feedback networks.

A second example deals with the noise performance of an indirect-feedback current amplifier as depicted — together with the relevant noise sources — in figure 3.11. The active part is assumed again to have nullor properties, while the transfer properties of the output stages are described by their transmission parameters.

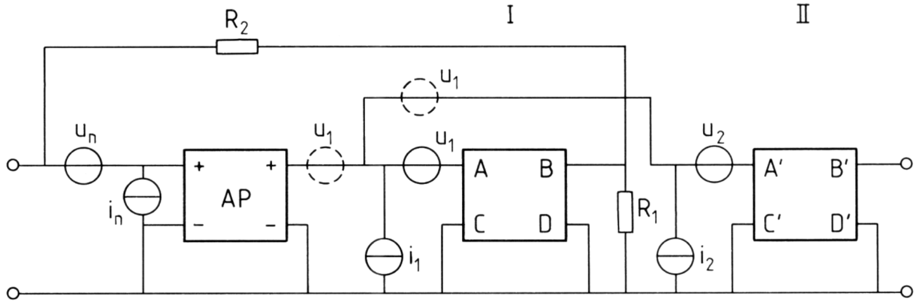


Figure 3.11 Indirect-feedback current amplifier with relevant noise sources. Thermal noise sources of the resistors are not shown.

The source u_n undergoes a transformation similar to that in the corresponding passive-feedback configuration. The thermal noise sources of R_2 and R_1 manifest themselves at the input in the same way as well. The input sources i_1 and i_2 of the output stages do not play any role because the parameters ζ and α of the active parts are zero. The sources u_1 and u_2, however, cause a short-circuit output current in stage II given by $i_o = (u_1 + \mathrm{u}_2)/B'$. When transformed back to the input, an equivalent current source i_{eqo} results with a value

$$i_{eqo} = i_o/\alpha_\infty = (u_1 + u_2)/B'\alpha_\infty$$

where α_∞ is the asymptotic-current-gain value of the complete amplifier. The noise contribution due to this source may be significant or even dominant for small values of α_∞, if the output stages are not carefully designed.

In section 3.5 we will pay attention to the simplest indirect-feedback current-amplifier configurations and show that these are inferior to the corresponding direct-feedback configurations, especially as a consequence of the influence of these noise sources u_1 and u_2.

In chapter 1 we have already paid some attention to the noise performance of an amplifier with indirect voltage sensing. We have shown there that the use of several series connected stages in the dummy input stage reduces the influence of the noise sources of this stage.

3.5 Comparison of the noise performance of single-device configurations

3.5.1 Introduction

One of the most important criteria for the selection of the proper input-stage configuration of an amplifier is the noise production of this stage. We will in this section compare the noise performance of the single-device configurations that were presented in figure 2.4. The comparison will be based on the performance of the CE configuration.

The noise of the stages is represented for this purpose by equivalent input noise sources, as indicated in figure 3.12. Because the input noise sources are correlated in general, this representation is not always most appropriate for the characterisation of noise performance. It is useful, however, for the above-described purpose.

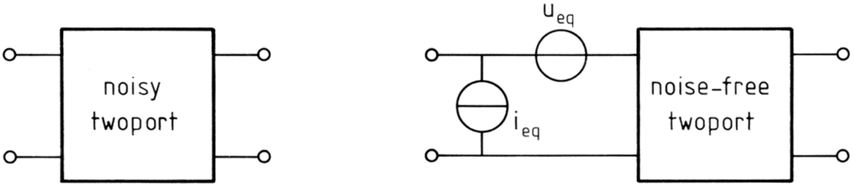


Figure 3.12 A noisy two-port represented as a noise-free two-port with two equivalent input noise sources.

In chapter 2 it was pointed out that eight basically different configurations with one active device can be distinguished (excluding configurations with transformers). Seven of them can be considered as feedback versions of the common-emitter stage (or in the case of an FET the common-source stage) . Two of these, i.e. the CB and the CC stages, have no impedances in the feedback network and can therefore be indicated as configurations with non-energic feedback. If their loop gain were infinite, their equivalent input noise sources would be equal to those of a CE stage. Because of the finite loop gain they will appear to be only slightly different. The noise performance of these stages is investigated first.

3.5.2 Noise of the non-energic-feedback stages

Figure 3.13a shows the CE stage with its input noise sources i_n and u_n. The CC stage is shown in figure 3.13b with the same noise sources. In order to obtain a representation with two equivalent noise sources at the input, the source i_n has to be transformed to the input. The transformation steps are shown in figure 3.14. Because of the small values of B and D (their approximate values are given in chapter 2), the sources are approximately equal to those of the CE stage, except at very high frequencies.

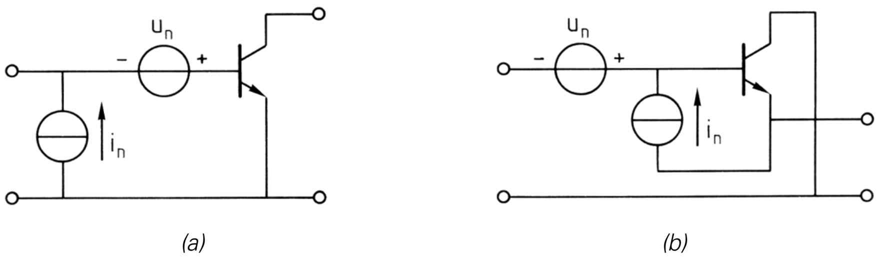


Figure 3.13. (a) CE stage with equivalent input noise sources i_n and u_n; (b) CC configuration with i_n and u_n.

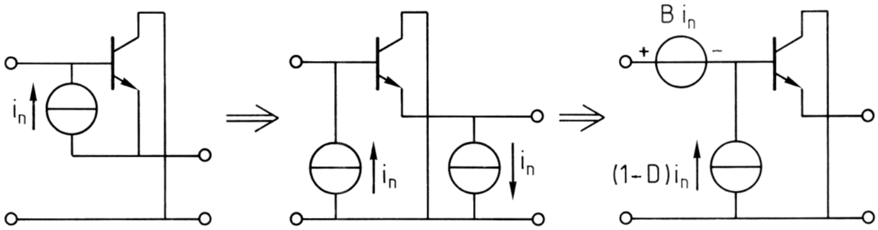


Figure 3.14 Transformation of i_n into equivalent input sources. B and D are transmission parameters of the CC configuration.

For the CB stage, a similar transformation can be performed, illustrated in figure 3.15, where A and C are transmission parameters of the CB configuration. Because the values of A and C are small, the equivalent input noise sources of the CB stage are nearly equal to those of the CE stage.

3.5.3 Noise of the impedance-feedback stages

The equivalent input noise sources of the other feedback configurations given in chapter 2 can be determined in a similar way. For the series stage and the shunt stage, the results are given in figure 3.16 and 3.17. The impedances Z_S are assumed to be noiseless. If they are not, the thermal noise sources can be taken into account just as with the corresponding feedback configurations of section 3.3. The transmission parameters given in the figures are related to the configurations involved (approximate values are given in chapter 2), while i_n and u_n are the equivalent input noise sources of the CE configuration.

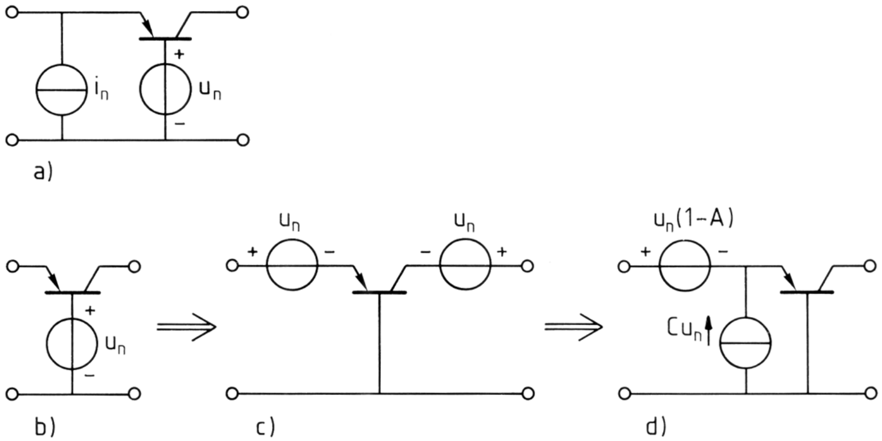


Figure 3.15 Transformation of noise sources in a CB stage.

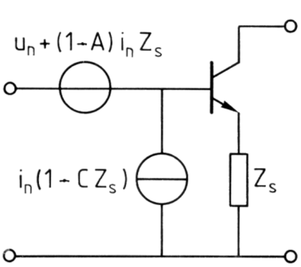


Figure 3.16 Equivalent input noise sources of the series stage.

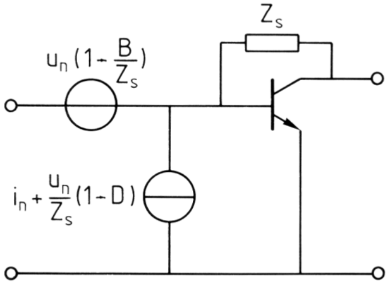


Figure 3.17 Equivalent input noise sources of the shunt stage.

The influence of the feedback impedances — including their thermal noise contributions — in the shunt and series stages on the noise performance manifests itself in a way very similar to that of the complete amplifiers as discussed in section 3.3. The finite values of the transfer parameters of the active part give use to differences of little significance. This may lead us to the conclusion that noise performance is influenced in the impedance-feedback cases (including those with more than one feedback loop) as if there were shunt or series impedances at the input. Consequently, noise performance is influenced in a generally adverse way.

Comparing now the noise performance of the various passive-feedback single device input stages, we may conclude that the CE stage together with the nonenergic feedback configurations (the CC stage and the CB stage) perform the best. As far as the noise production of the input stage itself is concerned they are the most preferable input stages.

3.5.4 Noise of indirect-feedback stages

It can easily be shown that the simple indirect-feedback configurations given in section 2.5 are inferior as input stages as a consequence of the noise contributions of the duplicate stages.

As an example, the current mirror is shown in figure 3.18 with the equivalent input noise sources of both transistors. An exact calculation of the equivalent input noise sources of this circuit is rather complicated, but these sources can be estimated quite easily by transforming the source u' to the input via the transmission parameters A and C of the input device. These parameters have the values unity and $1/B$ (CE stage parameter), respectively, so that approximately a total equivalent input voltage source $u + u'$ is found, while the input noise current source has a value $i_{eq} = i + i' + u'/B$. Especially the last term in this expression has a pernicious influence on the noise performance of the current mirror. Therefore, it is completely unsuitable as an amplifier input stage. Nevertheless, it has been proposed incidentally for this purpose [3].

Indirect-feedback stages with a current gain larger than unity have a less inferior noise performance.

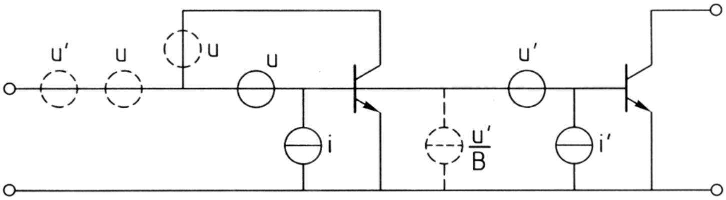


Figure 3.18 'Current mirror' with equivalent noise sources of both stages.

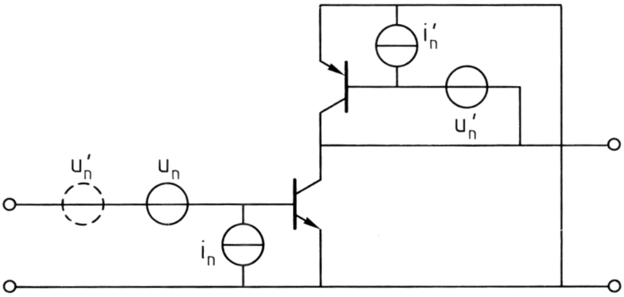


Figure 3.19 'Voltage mirror' with noise sources.

In the case of the simple configuration of figure 3.19 with indirect voltage comparison at the input (voltage mirror) it can easily be shown that u'_n can be transformed into a nearly equal source in series with u_n, thus doubling the spectrum of the input voltage source. The influence of i'_n will be negligible in practical cases. A smaller contribution of the dummy input stage to the input noise voltage is obtained if the gain is increased by using a number of series-connected stages, as discussed in chapter 1, section 1.5.2.

3.5.5 Noise of balanced stages

We will not deal here in detail with the noise performance of balanced stages such as the well-known differential pair. It will be clear from the general balanced configuration as depicted in figure 3.20 that the spectrum of the input noise voltage is doubled with respect to that of a single-sided stage, while the influence of the noise current sources depends on the signal-source characteristics. In general, noise performance of an amplifier with a differential input stage will therefore be somewhat inferior to that with an unbalanced configuration.

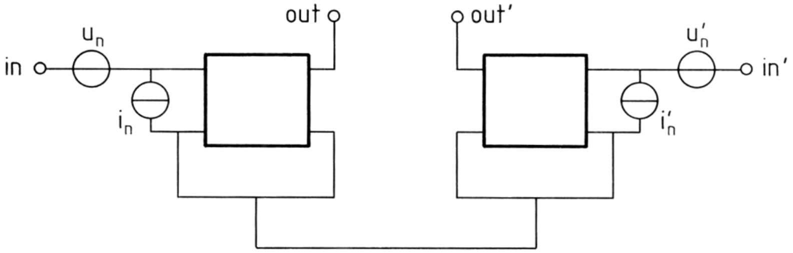


Figure 3.20 Balanced configuration with noise sources.

3.6 Considerations regarding the choice of the input configuration in relation to the noise contribution of a second stage

In the previous section we found that the CE, CB and CC configurations have nearly equal equivalent input noise sources. In the other single-device configurations the presence of series and shunt impedances leads to worse noise performance, even when the impedances are lossless and do not produce noise themselves. A justifiable choice of the input stage configuration cannot be made, however, before the influence of the second-stage noise is investigated. We will deal with second-stage noise contributions in this section.

The noise sources of the second stage of an amplifier, including the load impedance of the first stage, can always be represented as given in figure 3.21. The sources u and i can be described by their energy density spectra S_u and S_i, respectively; they will be supposed to be uncorrelated.

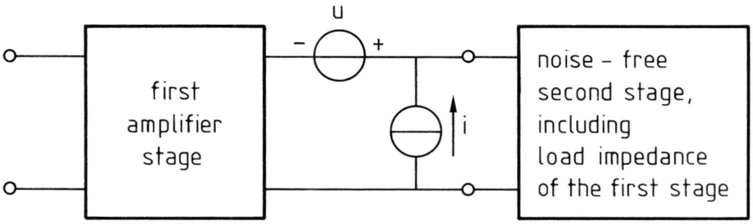


Figure 3.21 Representation of the noise sources of a second stage.

Formally the equivalent noise contribution from the second stage at the input of an amplifier can be calculated with Friis' formula [4] for the noise figure of a cascade of amplifier stages:

$$F_{tot} = F_i + \frac{F_2 - 1}{G_1} + \dots \frac{F_n{}^{-1}}{G_1 G_2 \dots G_{n-1}}$$

where F_i and G_i are the noise figure and the available power gain of the i-th two-port, respectively.

Quite a lot of calculation effort is required to find the noise contribution of the second stage, using this formula. Moreover, it does not reveal very directly the design criteria for the first stage. It merely states that — to have a small second-stage noise contribution — the available power gain of the first stage should be large, while the noise figure of the second stage should be small.

To find more operative design criteria for the first stage, we will determine the noise contribution of the second stage in a more direct way. The classification of single-device configurations, as given in chapter 2, will appear to be most useful for this determination.

Characterising again the transfer properties of the first amplifier stage by its transmission parameters A, B, C and D, the second-stage noise sources can easily be transformed into equivalent input sources, according to figures 3.3 and 3.4 in section 3.2. Supposing that the signal source is given as a voltage source with source resistance R_S, the total noise voltage contribution at the input, due to the second stage, can be found from figure 3.22.

The spectrum of the source u_{neq} is given by

$$S(u_{neq}) = S(u) \left| \frac{1}{\mu} + \frac{R_S}{\zeta} \right|^2 + S(i) \left| \frac{1}{\gamma} + \frac{R_S}{\alpha} \right|^2 \tag{3.2}$$

where μ, γ, ζ, and α are the transfer parameters of the first stage, being the reciprocal values of A, B, C and D, respectively, and $S(u)$ and $S(i)$ are the spectra of the second-stage noise sources.

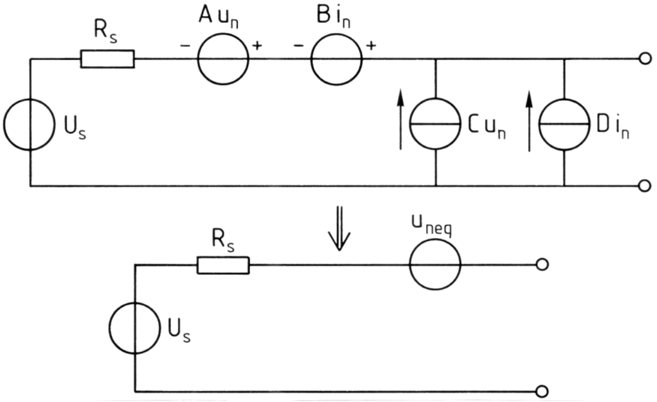


Figure 3.22 Transformation of second-stage equivalent noise sources into one equivalent noise voltage source at the input.

Expression (1) shows that all transfer parameters of the first stage occur in the denominators. In order to obtain a minimum noise contribution from the second stage, the values of all *transfer parameters* of the first stage should be *as large as possible*.

This is an important conclusion. When it is combined with the conclusion of the previous section, where we have seen that the CE, CB and CC stages perform the best as far as their own noise production is concerned, we now find that the CE stage (or the CS stage in the case of an FET) is the most preferable input stage of an amplifier.

Summarising the two reasons leading to a preference for a CE stage as the amplifier input stage:

1. Its noise performance can not be surpassed by that of other stages.
2. It has the largest transfer parameters so that the second-stage noise contribution is maximally reduced.

It should be emphasised that the *actual gain* of the first stage is not a criterion for the noise contribution of the second stage. The transformation of the second-stage noise sources is governed by the *gain parameters* of the first stage. In Friis' formula this is expressed by the presence of the *available* power gain.

When the application of a balanced input stage is imperative because of the frequency range to be covered, it is of course preferable to avoid local feedback in this stage as well. In the case where the active part of the negative-feedback amplifier must have a floating input port, and a single-device input stage has to be used (see section 2.6), local feedback in the input stage cannot always easily he avoided. It follows from the arguments above that in that case the reduction in the transfer parameter in question should be as small as possible. Therefore, when the most favourable input stage in such a case turns out to be a 'series stage' as depicted in figure 3.23 for a voltage amplifier, the feedback impedance Z_1 must be as small as possible for two reasons:

1. The feedback impedance causes an increase in the influence of the noise current source.
2. The reduction in the transconductance (the transfer parameter affected by feedback) causes an enhancement of the noise contribution of the second stage.

If the feedback impedance itself produces noise, there is a third reason to keep at least its real part small.

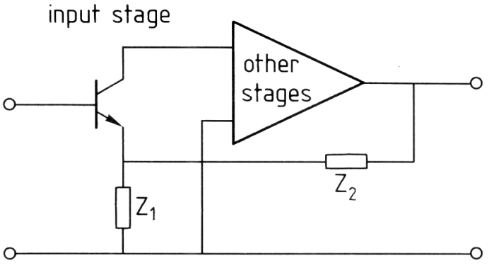


Figure 3.23 Voltage amplifier with a 'series-stage' input configuration.

Summarising, we can conclude that the selection of the input-stage configuration is concerned in the first place with the noise production of that stage itself, while in the second place the noise contribution of the second stage must be taken into account. CE and CS stages offer the best performance in this respect because they have the largest transfer parameters (and consequently, the largest available power gain).

In the following chapters it will appear that there are no decisive reasons to prefer other types of input stages for the optimisations with respect to other quality aspects. The selection of the input-stage configuration can therefore be predominantly based on noise considerations.

The second stage has to be designed so that its noise contribution has a reasonably small value. Basically, the use of a coupling transformer with an optimised turns ratio would be required between the stages in order to minimise the second-stage noise contribution. The need for minimisation, however, is frequently not very urgent if all transfer parameters of the first stage are large.

The influence of the second-stage noise sources can be estimated quickly and taken into account when necessary, by making use of a representation with transmission parameters of the first stage and of the approximate values of these parameters as given in table 2.3.

3.7 Considerations regarding the selection of the type of input device and its bias conditions

3.7.1 Introduction

In the foregoing sections it was pointed out that the best noise performance can be achieved with a non-feedback stage at the input (i.e. a CE or a CS configuration) of an amplifier, for two reasons:

1. The signal-to-noise ratio obtainable with non-feedback configurations cannot be surpassed by that of local-feedback stages.
2. One or more transfer parameters of the local-feedback configurations are smaller, resulting in larger noise contributions of the second stage.

In this section we will present criteria for the selection of the type of input device and its bias conditions with different types of source impedances. We will assume in the following that the overall-feedback network and the second amplifier stage do not contribute to the amplifier noise. Should they happen to, they can be brought into account by the methods given in the previous sections. Furthermore, we will deal exclusively with non-feedback input stages.

The selection criteria for an input active device have to be derived from its noise properties, which will be discussed first.

3.7.2 Noise models of active devices

In order to formulate the desired criteria and to develop an optimisation procedure we will base ourselves on the simple noise models as presented in the figures 3.24 and 3.25.

(i) Noise sources in the bipolar transistor

The noise sources in the bipolar-transistor model of figure 3.24 can be considered as uncorrelated at all but very high frequencies.

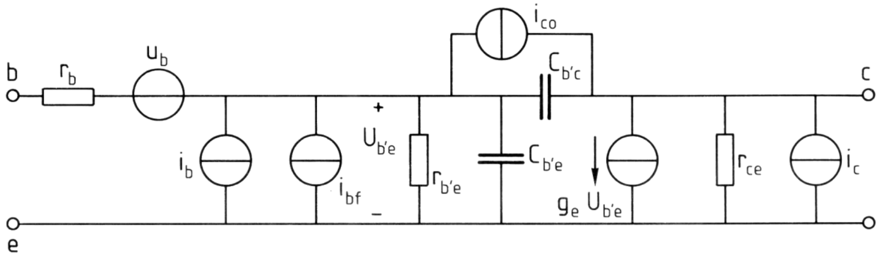


Figure 3.24 Noise model of a bipolar transistor.

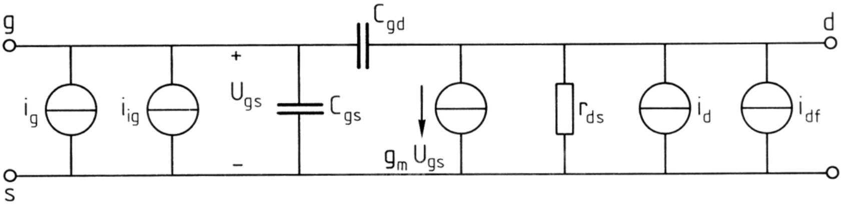


Figure 3.25 Noise model of a field-effect transistor.

The shot noise (i_c) of the collector current has a power-density spectrum $S(i_c) = 2qI_C$. In the following we will frequently denote $S(i_c)$ as a thermal equivalent by using the relation $r_e = kT/qI_C$, which yields $S(i_c) = 4kT/2r_e$.

The shot noise i_b of the base current has a spectrum $S(i_b) = 2qI_B = 4kT/2h_{FE}r_e$. The thermal equivalent expression for this spectrum is obtained via the same relation between r_e and I_C as above.

The excess or $1/f$ noise source (i_{bf}), working in parallel with the base current source, can be characterised by a spectrum $S(i_{bf}) = KI_{\alpha}^{\beta}/f$, where K is a constant dependent on the manufacturing process and α varies between the values 1 and 2. Alternatively, we can write $S(i_{bf}) = 2qI_Bf_\ell/f$, or $S(i_{bf}) = 4kTf_\ell/2h_{FE}r_ef$, where f_ℓ is called the excess-noise corner frequency for which $S(i_b) = S(i_{bf})$. The frequency f_ℓ, which can be regarded as a figure of merit of the transistor (unfortunately rarely specified by the manufacturer), has a value between some Hz and some tens of kHz [5], [6].

The base resistance is assumed to produce thermal noise: $S(u_b) = 4kTr_b$.

The shot-noise source i_{co}, which is related to the collector-base leakage current, has a spectrum $S(i_{co}) = 2qI_{CO}$. It can be ignored in silicon transistors unless bias currents become extremely small.

(ii) Noise sources in the field-effect transistor

In the field-effect transistor, the source i_g represents the shot noise of the gate leakage current (junction FET). It has a spectrum $S(i_g) = 2qI_G$.

The thermal noise of the channel is represented by two correlated sources i_d and i_{ig} with spectra [7]:

$$S(i_d) = 4kTcg_m \qquad \text{and} \qquad S(i_{ig}) = 4kT\frac{\omega^2 C_{gs}^2}{3g_m}$$

Practical values of c range from one to two. The source i_{ig} is generally known as the induced gate noise. It is shown in the appendix that the induced gate noise can be neglected without making a large error in the case of wide-band amplifiers. We will ignore this source in the following sections.

The most important excess noise source i_{df} has a spectrum which can be written as

$S(i_{df}) = 4kTcg_m f_\ell/f$, where f_ℓ is the frequency in which $S(i_{df}) = S(i_d)$. For junction FET's f has a value between 1 kHz and 100 kHz.

In MOSFET's f_ℓ is usually much larger. Only in wide-band amplifiers, where for example frequencies above 1 MHz are amplified, may the MOSFET be a useful input device.

(iii) Comparison of bipolar-transistor and FET behaviour

Comparing the noise models of the bipolar transistor and the FET, we find some similarities as well as some differences. First, the bipolar transistor has a thermal input noise voltage source due to the base spreading resistance not present in the FET. Second, the input noise current source of the bipolar transistor is much larger than that of the FET because in the first case the input bias current is larger. Third, the excess noise in the bipolar transistor is present in the input bias current, whereas, in the case of an FET it is found in the output current source.

The above differences may lead in a certain application to a preference of one type of device over the other, as will appear in the subsequent sections.

3.7.3 Transformation of noise sources to the input

In order to compare the performances of different devices, it will be useful to transform all noise sources to the input of the stage. A transformation into the noise model with two equivalent input sources as used in section 3.3 is considered unwieldy because it leads to correlated sources at the input. Moreover, this model requires yet an additional transformation for the calculation of the contribution of the amplifier noise to the signal and noise from the signal source.

Because the signal source is usually represented by a Thévenin or Norton equivalent circuit, a transformation of all noise sources into one voltage source in series with, or one current source in parallel with the signal source is preferred. Figure 3.26 shows these two situations for a bipolar-transistor input stage, where the spectra of u_{eq} and i_{eq}, respectively, have to be calculated to find the signal-to-noise ratios or noise figures.

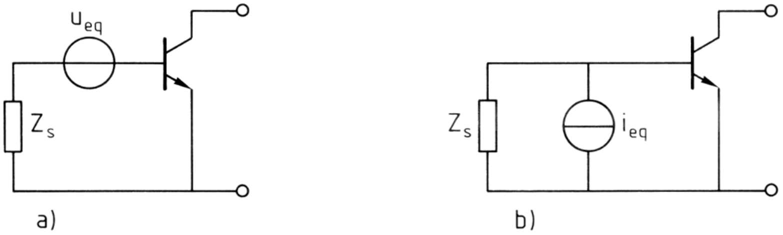


Figure 3.26 Desired models for Thévenin and Norton representations of the signal source.

As an example we will perform the calculation of $S(u_{eq})$ for a bipolar transistor with a source impedance Z_s connected to it. The equivalent circuit is depicted in figure 3.27. Using the transmission-parameter representation for the transistor, an initial transformation step leads to the circuit of figure 3.28 where the source i_c is transformed into two sources Bi_c and Di_c. Because these are correlated, their signs are preserved for the time being. The parameters B and D of the intrinsic transistor have the values

$$B = \frac{1}{\gamma} = -r_e = -\frac{kT}{qI_C} \qquad \text{and} \qquad D = \frac{1}{\alpha} = -\left(\frac{1}{\alpha_{eo}} + \frac{j\omega}{\omega_T}\right)$$

where, referring to figure 3.27,

$$\alpha_{eo} = g_e r_{b'e} \qquad \text{and} \qquad \omega_T = \frac{g_e}{C_{b'e} + C_{b'c}}$$

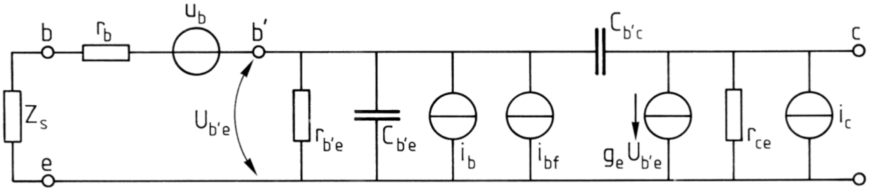


Figure 3.27 Bipolar-transistor equivalent circuit with noise sources and source impedance Z_s.

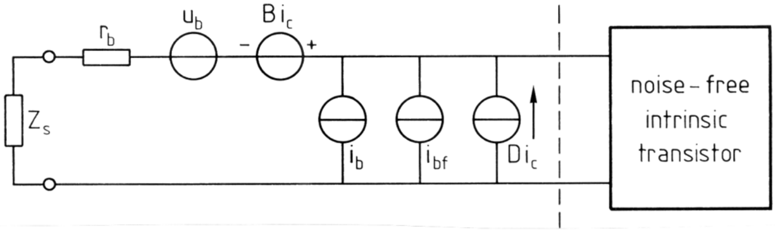


Figure 3.28 Transformation of the source i_c to the input.

The result of a second transformation leading to the desired single noise source in series with the signal source is given in figure 3.29. The voltage source in is given by:

$$u_{eq} = u_b + i_c(B + D(r_b + Z_s)) + (i_b + i_{bf})(r_b + Z_s) \tag{3.2}$$

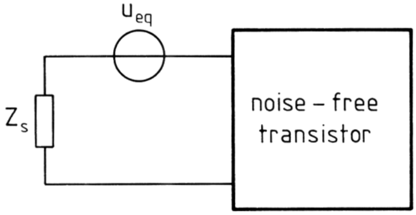


Figure 3.29 Result of a second transformation step.

The sources u_b, i_c, i_b and i_{bf} are uncorrelated, which renders calculation of the spectrum $S(u_{eq})$ simple. The result with the above values for the parameters B and D and the expressions for the spectra given earlier is:

$$S(u_{eq}) = 4kT\left[r_b + \frac{|r_b + Z_s|^2}{2h_{FE}r_e}\left(1 + \frac{f_\ell}{f}\right) + \frac{1}{2r_e}\left| r_e + \left(\frac{1}{\alpha_{eo}} + \frac{j\omega}{\omega_T}\right)(r_b + Z_s)\right|^2\right] \tag{3.3}$$

To find the total input noise voltage, the spectrum of the noise voltage of the signal source has to be added to $S(u_{eq})$:

$$S(u_{tot}) = 4kTRe(Z_s) + S(u_{eq}) \tag{3.4}$$

A reduction of all noise sources to one equivalent noise *current* source at the input in the case where the signal source is represented as a Norton equivalent circuit can be similarly performed and leads to:

$$S(i_{eq}) = \frac{S(u_{eq})}{|Z_s|^2} \tag{3.5}$$

where $S(u_{eq})$ is given by expression (3.3).
The spectrum of the total input noise current is found again by adding the noise of the signal source:

$$S(i_{tot}) = 4kTRe(Y_s) + S(i_{eq}). \tag{3.6}$$

In the case of a field-effect transistor the transformations leading to a single noise source at the input for given source impedances are very similar.
The spectrum of u_{eq} is given in this case by:

$$S(u_{eq}) = 4kTcg_m\left(1 + \frac{f_\ell}{f}\right)\left|\frac{1}{g_m} + Z_s\frac{j\omega}{\omega_T}\right|^2 + 2qI_G|Z_s|^2 \tag{3.7}$$

The spectrum of i_{eq} is related to $S(u_{eq})$ of expression (3.7) in the same way as it was in the case of the bipolar transistor (expression (3.5)).
The expressions for the spectra found here are the basis for that part of the design, in which the proper input device has to be found for a given signal source. They

can be used for calculating signal-to-noise ratios, spot-noise figures, etc.

As a reminder it is noted here that the influence of an overall feedback network and of the second stage on the above input spectra can, if necessary, be taken into account according to the methods discussed in sections 3.4 and 3.6.

3.8 Design procedure for optimum noise match of the amplifier to the source

3.8.1 Introduction

Now that we have derived the expressions of the noise spectra at the amplifier input, all ingredients for the optimisation of the signal-to-noise ratio or the realisation of a so-called noise match are available. Two different ways of achieving a wide-band noise match can be distinguished. The first procedure matches the signal-source impedance to the amplifier with the aid of a coupling transformer. In this case the reflected source impedance enters in the expressions for $S(u_{eq})$ or $S(i_{eq})$, and the noise contribution of the amplifier can be minimised by finding the optimum turns ratio of the transformer. This approach is considered unsuitable because the use of a transformer must usually be excluded for practical reasons.

An alternative procedure where the input stage is adapted to the source impedance should be followed when a coupling transformer cannot be tolerated. This alternative approach for maximising the signal-to-noise ratio will be emphasised here and is described below.

Depending on the type of signal information produced by the signal source, the total noise voltage, current or charge is calculated by integrating the spectrum over the relevant frequency region, or — if the weighted signal-to-noise ratio has to be maximised — the spectrum is multiplied first by the weighting function and the integration is performed afterwards. The capability of a bipolar-transistor type to achieve an optimum signal-to-noise ratio is judged on the grounds of the influence of the various parameters (viz. r_b, h_{FE}, α_{eo}, r_e, ω_T and K) on the total noise.

The best device cannot be found analytically, because the relations between various parameters are known only qualitatively and are dependent on the manufacturing process. A further complication is brought about by the fact that manufacturers give poor information concerning the noise performance of their devices. Values of r_b and a specification of excess-noise behaviour are rarely given.

Because the parameter r_e occurs in the numerator as well as in the denominator of some terms of the expression for the total noise, as can be seen from (3.3) a minimum in the total noise can be approached by varying the bias current, provided that the other parameters are less dependent on the bias current. This minimum can be calculated for various transistor types if the relevant characteristics are known, and the best type can be selected or manufactured [8], [9].

A similar procedure can be followed with a field-effect transistor. If ω_T is considered to be a constant one can find a maximum SNR by using a device with an optimum g_m value, or by varying the number of paralleled input devices [10], [11], [12].

The most suitable input device can be selected on the grounds of a comparison between the optimisation results for the bipolar and field-effect transistor.

For some types of source impedances the noise performance of an amplifier with either a bipolar transistor or a field-effect transistor in the input stage will be investigated in some detail in the subsequent sections. We will deal with predominantly resistive, capacitive and inductive signal sources and discuss the minimisation of the amplifier noise contribution. Of course it is possible to extend the treatment to more complicated source impedances, but the conclusions which are valid for these three types of source impedances still hold in certain frequency regions for more complicated source impedances.

It is not the purpose of the following sections to give detailed design considerations for all types of sources. In order to provide an impression of the design procedure as outlined before we will restrict ourselves to the consideration of wide-band amplifiers, where the condition that $f_2 \gg f_1$ is met, f_2 and f_1 being the high- and low-frequency band edges, respectively. In that case the $1/f$ noise sources can be assumed to make an insignificant contribution to the unweighted total noise. The influence of the $1/f$ noise sources on the spectrum, however, can be troublesome and therefore it will not be disregarded completely.

3.8.2 Noise matching with real source impedances

For the minimisation of the amplifier noise contribution in the case of a real source impedance it is immaterial whether the total noise voltage or current is considered, because both spectra have the same frequency dependency. We will investigate and compare the noise performance of the bipolar and field-effect transistor in this section.

(i) Bipolar-transistor input stage

Dealing only with the spectrum of the voltage source in the case of a bipolar-transistor input stage, we can write:

$$S(u_{tot}) = 4kTR_s + S(u_{eq}) \tag{3.8}$$

where $S(u_{eq})$ is given by (3.3), and it is assumed that the source impedance R_s produces thermal noise.

The excess-noise spectrum plays a significant cant role only in the low-frequency region. In this region $\alpha_e(\omega)$ equals α_{eo} and, ignoring some small contributions, the low-frequency part of the spectrum $S(u_{tot})$ can be written as:

$$S(u_{tot})_{(l.f)} = 4kT(R_s + r_b + \tfrac{1}{2}\, r_e) + \frac{4kT}{2h_{FE}r_e}\left(1 + \frac{f_\ell}{f}\right)(r_b + R_s)^2 \tag{3.9}$$

The asymptotes of the spectrum described by (3.9) are drawn in figure 3.30. The contributions of the various terms are chosen arbitrarily.

For frequencies lower than f_ℓ, the $1/f$ noise contribution dominates. With a given source resistance, the value of strongly depends on the type of transistor and its bias current. As the source resistance gets smaller, the relative contribution of the excess-noise source likewise decreases.

We will now determine the total noise voltage, but only for the *wide-band* amplifier case where the excess-noise contribution can be neglected. The spectrum of the total noise voltage can then be approximated as:

$$S(u_{tot}) = 4kT(R_s + r_b + \tfrac{1}{2}\, r_e) + \frac{4kT}{2r_e}\left(\frac{1}{h_{FE}} + \frac{\omega^2}{\omega_T^2}\right)(r_b + R_s)^2 \tag{3.10}$$

The total noise voltage is found by integrating the spectrum over the relevant frequency band. Assuming $f_2 \gg f_1$:

$$\overline{u_{tot}^2} = \int_{f_1}^{f_2} S(u_{tot})\cdot \mathrm{d}f = 4kTf_2\left[(R_s + r_b + \tfrac{1}{2}\, r_e) + \frac{r_b + R_s}{2r_e}\left(\frac{1}{h_{FE}} + \frac{1}{3}\frac{\omega_2^2}{\omega_T^2}\right)\right] \tag{3.11}$$

The term with ω_T in (3.11) gives a significant contribution if $\omega^2 \geq 3\omega_T^2/h_{FE}$. The minimum of $\overline{u_{tot}^2}$ can be determined by varying r_e. For the optimum value of r_e it is found that:

$$r_{eopt}^2 = \left(\frac{1}{h_{FE}} + \frac{1}{3}\frac{\omega_2^2}{\omega_T^2}\right)(r_b + R_s)^2 \tag{3.12}$$

and for this value of r_e expression (3.10) becomes:

$$\overline{u_{tot}^2} = 4kT(R_s + r_b + r_{eopt})f_2$$

For different types of transistors, r_{eopt} can be calculated from (3.12). This calculation is of an iterative type, because r_b, ω_T and h_{FE} also depend on the bias current. Large values of h_{FE} and ω_T and a low value of r_b are advantageous. At low values of the source resistance, r_b can have a predominant influence on the total noise.

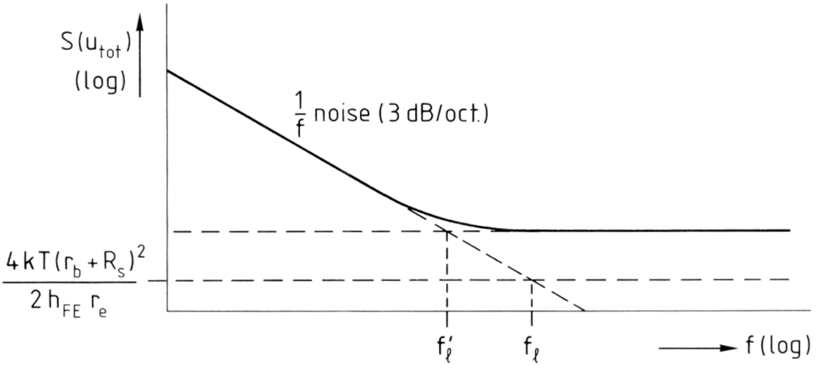


Figure 3.30 *Spectrum of the total noise voltage for a bipolar-transistor input stage and a real source impedance at relatively low frequencies.*

For a given transistor type the maximum values of h_{FE} and ω_T and a relatively small value of r_b are obtained with approximately the same value of the bias current. Therefore, with low source resistances a transistor type should be selected where ω_T and h_{FE} reach their maximum at a large bias current, i.e. a transistor with a large emitter area [9], [13]. With high source resistances small transistor types are favoured. The value of r_b is less critical in that case. As the optimum bias current becomes smaller, it will be more difficult to find transistor types with large values of ω_T and h_{FE}. The tendency of integrated circuits technology toward smaller device dimensions is advantageous in this respect.

The consideration of the influence of ω_T on noise performance is meaningful only in the case of amplifiers with a very large bandwidth. Frequently, a more simple expression for r_{eopt} can be used, viz.:

$$r_{eopt} = \frac{r_b + R_s}{\sqrt{h_{FE}}}$$

Expression (3.13) for $\overline{u_{tot}^2}$ can then be written in an alternative form:

$$\overline{u_{totmin}^2} = 4kT(r_b + R_s)\left(1 + \frac{1}{\sqrt{h_{FE}}}\right) \cdot f_2$$

The minimum value of the noise figure F_{min} in this case is given by:

$$F_{min} = \left(1 + \frac{1}{\sqrt{h_{FE}}}\right)\left(1 + \frac{r_b}{R_s}\right) \tag{3.14}$$

From (3.14) it is seen that a small noise figure can be obtained for source resistances large with respect to r_b, provided that h_{FE} is not reduced too much in the case of low bias currents. When the reduction of h_{FE} is too large, the field-effect transistor can be an alternative.

(ii) FET input stage

The low-frequency part of the spectrum in the case of an FET is given by:

$$S(u_{tot}) = 4kT\left[R_5 + \frac{c}{g_m}\left(1 + \frac{f_\ell}{f}\right)\right] + 2qI_G R_s^2$$

In contrast to the bipolar transistor, the relative excess-noise contribution increases as the source resistance gets smaller.

For the total noise voltage at the input of an amplifier with an FET input stage we find with the aid of expression (3.7):

$$\overline{u_{tot}^2} = 4kT\left[\left(R_s + \frac{c}{g_m}\right) + cg_m \frac{1}{3}\frac{\omega_2^2}{\omega_T^2} R_s^2\right] f_2 + 2qI_C R_2^2 f_2 \qquad (3.15)$$

Here also the $1/f$ contribution is disregarded, and it is assumed that $f_2 \gg f_1$.

In the case of very wide-band amplifiers where the second term in (3.15) cannot be neglected, an optimum value of can be found, assuming that ω_T is a constant. It is possible to use several FET's in parallel, resulting in a 'device' with the same value of ω_T as one FET and with an enlarged g_m [12]. Besides, the state of the art shows that different types of FET's can have the same value of ω_T for different values of g_m, so that it will frequently be possible to find the proper FET for a certain application.

Determination of the minimum of $\overline{u_{tot}^2}$ by varying g_m yields:

$$g_{mopt}^2 = \frac{\omega_T^2}{\omega_2^2}\frac{3}{R_s^2} \qquad (3.16)$$

the minimum value of $\overline{u_{tot}^2}$ is given by:

$$\overline{u_{tot}^2} = 4kT\left(R_s + \frac{2c}{g_{mopt}} + \frac{R_s^2}{R_{GC}}\right) f_2 \qquad (3.17)$$

where R_{GC} is found from the definition of the equivalent thermal noise:

$$2qI_G = \frac{4kT}{R_{GC}}$$

so that:

$$R_{GC} = \frac{2kT}{q}\cdot\frac{1}{I_G} \approx 50\cdot 10^{-3}\cdot I_G^{-1}$$

at room temperature.

The third term in (3.17) will be negligible except for very large source impedances or high temperatures. As a matter of fact, the shot noise of the gate leakage current will usually be lower than the noise of a gate leakage resistor, which will frequently be needed for biasing purposes. From (3.16) and (3.17) it follows that a large and frequently not practical value of g_{mopt} is found, when $\omega_2 \ll \omega_T$ and the source resistance is small.

(iii) Comparison of bipolar-transistor and FET performance.

Comparing the low-frequency situation of an FET stage with that of a bipolar-transistor input stage, but ignoring $1/f$ noise, we find minimum values for the noise contribution in both cases. With the FET the minimum noise contribution of the amplifier is given by:

$$S(u_{eq}) = 4kT\frac{c}{g_{mmax}}$$

and with the transistor by:

$$S(u_{eq}) = 4kTr_b$$

These limiting values may have the same order of magnitude for the devices available nowadays. In both cases large bias currents are necessary. The bipolar transistor, however, will generally be preferred because its noise contribution will approach the minimum value with much a smaller bias current.

When the bandwidth is so large that in the expression for r_{eopt} (3.12) the term with ω_T dominates, the devices can be approximately equivalent with respect to their noise performance, provided that devices with equal transit frequencies are available.

With relatively a small bandwidth and large source resistance, the noise performance of the FET is superior.

If the use of a coupling transformer at the input can be tolerated, a noise match can be realised by selecting the proper turns ratio provided that a high quality transformer can be practically realised. In that case the FET is preferred as an input device if the bandwidth is relatively small ($\omega_2 \gg \omega_T$).

For a very large bandwidth the FET and bipolar transistor are nearly equivalent when their transit frequencies ω_T are equal. The use of a transformer makes sense only when a better signal-to-noise ratio can be realised, or when on the grounds of other considerations a bias current is chosen which does not — unless the signal-source impedance level is transformed — lead to an optimum noise match.

3.8.3 Noise matching with capacitive source impedances

In signal sources with capacitive source impedances either the current, the voltage or the charge may be the best representation of the primary signal. From the expression of the spectrum of, for example, the equivalent current source, the spectra of the equivalent voltage or charge source can easily be obtained. The total noise current, voltage or charge can then be calculated from these expressions. The relevant quantity can be minimised by following a procedure similar to that of the previous section. As an example, we will deal here with the equivalent current source exclusively.

(i) Bipolar-transistor input stage

By ignoring the $1/f$ noise — because its influence is noticeable only in a small part of the frequency band — together with some other small contributions, the spectrum of this current source in the case of a bipolar transistor can be written as:

$$S(i_{tot}) = 4kT\Big[\Big(r_b + \frac{r_e}{2}\Big)\omega^2 C_s^2 + \frac{\omega}{\omega_T}\omega C_s + \frac{1}{r_e}\Big(\frac{1}{h_{FE}} + \frac{\omega^2}{\omega_T^2}\Big)(1 + \omega^2 C_s^2 r_b^2)\Big] \tag{3.18}$$

Integration of $S(i_{tot})$ over the frequency band f_1 - f_2, with $f_2 > f_1$, yields:

$$\overline{i_{tot}^2} = 4kTF_2\Big\{\frac{1}{3}\Big(r_b + \frac{r_e}{2}\Big)\omega_1^2 C_s^2 + \frac{1}{3}\frac{\omega_1^2 C_s^2}{\omega_T C_s} + \frac{1}{2h_{FE} r_e}\Big(1 + \frac{1}{3}\omega_2^2 r_b^2 C_s^2\Big) + \frac{\omega_2^2}{2r_e\omega_T^2}\Big(\frac{1}{3} + \frac{1}{5}\omega_2^2 C_s^2 r_b^2\Big)\Big\} \tag{3.19}$$

Determination of the minimum of (3.19) by varying r_e yields:

$$r_{eopt}^2 = \frac{1}{h_{FE}}\Big(r_b^2 + \frac{3}{\omega_2^2 C_s^2}\Big) + \frac{\omega_2^2}{\omega_T^2}\Big(\frac{3}{5}r_b^2 + \frac{1}{\omega_2^2 C_s^2}\Big) \tag{3.20}$$

With this value of r_e, (3.19) can be written as:

$$\overline{i_{tot}^2} = 4kT\frac{1}{3}f_2\omega_2^2 C_s^2\Big\{r_b + r_{eopt} + \frac{1}{\omega_T C_s}\Big\} \tag{3.21}$$

Expression (3.20) gives a criterion for the optimisation of the unweighted signal-to-noise ratio in wide-band amplifiers.

Though the $1/f$ contribution was disregarded, it is nevertheless meaningful — for the sake of a justifiable comparison between bipolar transistor and FET — to consider its influence on the spectrum $S(i_{tot})$. Figure 3.31 shows this spectrum. It is supposed that the spectra $4kT/2h_{FE}r_e$ and kI_B^{α}/f dominate in the low-frequency region. If the bias current is chosen so that $r_e = r_{eopt}$, the shaded area in figure 3.31 is minimised.

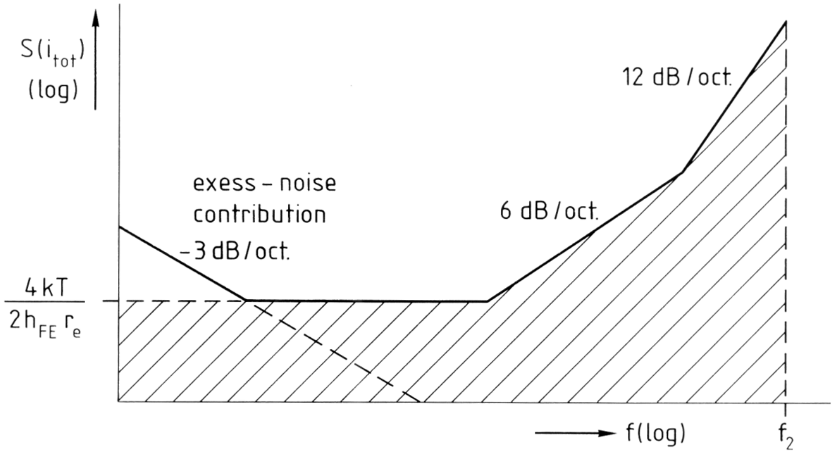


Figure 3.31. Spectrum $S(i_{tot})$ with capacitive source impedance for a bipolar transistor as an input device.

(ii) FET input stage

For the spectrum of the noise current source in the case of a junction FET it is found that:

$$S(i_{tot}) = 4kT\Big\{cg_m\left(1+\frac{f_\ell}{f}\right)\left(\frac{\omega^2}{\omega_T^2}+\frac{2\omega C_s}{g_m}\,\frac{\omega}{\omega_T}+\frac{\omega^2 C_s^2}{g_m^2}\right)+\frac{1}{R_{GC}}\Big\} \qquad (3.22)$$

In the low-frequency part of the spectrum the term $4kT/R_{CG}$, or rather a term due to the gate-bias resistor (not given in (3.22)), will generally dominate.

The $1/f$ noise dominates only in the low-frequency region in the first term in (3.22). Therefore, it frequently does not make any significant contribution in the spectrum, and thus leads to a performance essentially different from that of the bipolar transistor. The spectrum is drawn in figure 3.32. In the same way as with the bipolar transistor the minimum of $\overline{i_{tot}^2}$ can be determined by varying g_m. For wide-band amplifiers where $f_2 \gg f_1$, the result is:

$$g_{mopt}^2 = \omega_T^2 C_s^2 \qquad (3.23)$$

For this value of g_m, $\overline{i_{tot}{}^2}$ can be written as:

$$\overline{i_{tot}{}^2} = 4kT\left\{c\,\frac{4}{3}\,\frac{\omega_2^2 C_s^2}{g_{mopt}}+\frac{1}{R_{GC}}\right\}f_2 \qquad (3.24)$$

Condition (3.23) can (with $\omega_T = g_m/C_{iss}$) alternatively be written as:

$$C_{issopt} = C_s \qquad (3.25)$$

From (3.25) it can be seen that it is possible only in a restricted region of source impedances to find a junction FET for optimum noise matching. Large values of the source capacitance lead to values of the input capacitance of the FET, corresponding with possibly unpractically large values of g_m (ω_T is assumed to have the same value for different types of FET's). Very small values of C_s require possibly not available values of the input capacitance C_{iss}.

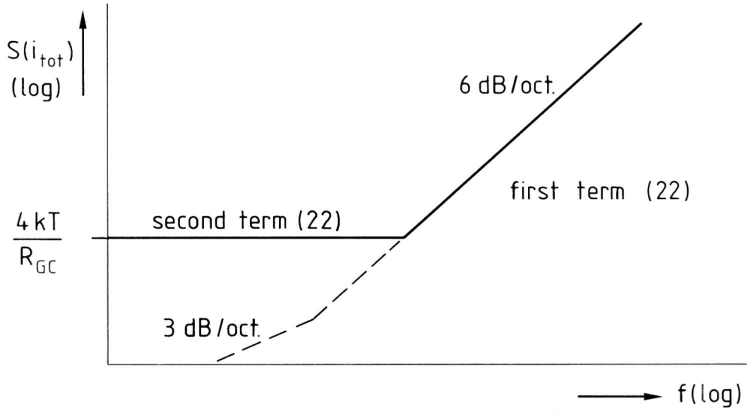


Figure 3.32. Spectrum $S(i_{tot})$ with a capacitive source impedance for an PET as an input device.

(iii) Comparison of bipolar-transistor and PET performance

With a bipolar transistor it follows from (3.20) that large values of C_s and ω_2, while $\omega_2 \ll \omega_T$, lead to small values of r_{eopt} ($r_{eopt} \approx r_b/\sqrt{h_{FE}}$). The minimum noise contribution of the input device in that case is determined by the value of r_b. A junction FET with a large value of and an optimum biased bipolar transistor are nearly equivalent in this situation.

With low values of ω_2 or C_s the junction FET is superior as an input device [11], [12], [14]. If an approximately non-energic coupling transformer at the input can be used, the FET is always the best input device provided that the proper turns ratio can be chosen.

3.8.4 Noise matching with inductive source impedance

The frequency region in which the inductive part of a source impedance dominates will generally be small because of the less ideal coil characteristics (series resistance, shunt capacitance). Therefore, there seems to be no reason to evaluate the integrated spectrum of the equivalent input noise source for the very wide-band amplifier case. The noise behaviour of the transistor and FET will be investigated with respect to the total noise voltage in the low-frequency region only.

(i) Bipolar-transistor input stage

By disregarding the input noise-current contribution $i_c D(r_b + Z_s)$ (expression 2), which can be allowed, provided that $\omega \ll \omega_T$ and because $h_{FE} \ll \alpha_{eo}^2$, the low frequency part of the spectrum $S(u_{tot})$ can be written as:

$$S(u_{tot}) = 4kT \left\{ r_b + \frac{1}{2} r_e + \frac{1}{2h_{FE} r_e} (r_b + \omega^2 L_s^2)\left(1 + \frac{f_\ell}{f}\right) \right\} \tag{3.26}$$

In practice, the shunt capacitance of the source inductance will restrict the frequency range over which the low-frequency approximation holds. The resulting low-frequency spectrum can be integrated over the relevant frequency band f_1 - f_2, and the integral can be minimised by varying r_e The contribution of the $1/f$ noise source to the total noise will be small compared to the contribution of the other terms. Integration of the low-frequency part of the spectrum then yields:

$$\overline{u_{tot}^2} = 4kTf_2 \left\{ r_b + \frac{1}{2} r_e + \frac{r_b^2}{2h_{FE} r_e} + \frac{\frac{1}{3} \omega_2^2 L_s^2}{2h_{FE} r_e} \right\} \tag{3.27}$$

The minimum of (3.26) is found for:

$$r_{eopt}^2 = \frac{r_b^2 + \frac{1}{3} \omega_2^2 L_s^2}{h_{FE}} \tag{3.28}$$

For this value of r_e, (3.27) can be rewritten as:

$$\overline{u_{tot}^2} = 4kTf_2 (r_b + r_{eopt}) \tag{3.29}$$

(ii) PET input stage

With the junction FET, the low-frequency part of the spectrum of the equivalent voltage source is given by:

$$S(u_{tot}) = 4kT \frac{c}{g_m} \left(1 + \frac{f_\ell}{f}\right) + 2qI_G \omega^2 L_s^2 \tag{3.30}$$

In wide-band amplifiers the influence of the $1/f$ noise on the total noise voltage can usually be neglected again. Integration of the spectrum between the boundaries f_1 and f_2 then yields:

$$\overline{u_{tot}^2} = 4kT \frac{c}{g_m} f_2 + \frac{2}{3} qI_G \omega_2^2 L_s^2 f_2 \tag{3.31}$$

To keep the noise voltage small, g_m should have a large value.

(iii) Comparison of bipolar-transistor and PET performance

Comparison of expressions (3.29) and (3.31) shows that the junction FET can be a favourable choice as an input device in the case of an inductive signal source, however, usually at the expense of a large bias current.

In the frequency region where the $1/f$ noise contribution is important, there is, as with capacitive sources, an essential difference between the noise performance of an FET and a bipolar-transistor input stage. This difference is best shown by a plot of the low-frequency part of the contribution to the spectrum. Figures 3.33 and 3.34 show these contributions and the resulting spectra for the transistor and the FET, respectively.

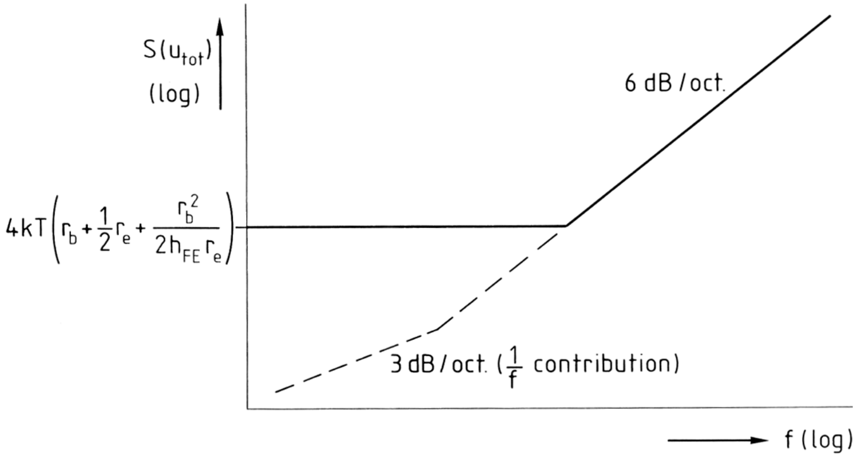


Figure 3.33. Spectrum of $S(u_{tot})$ (bipolar transistor) with inductive source impedance. The 1/f contribution is masked by the frequency-independent contributions.

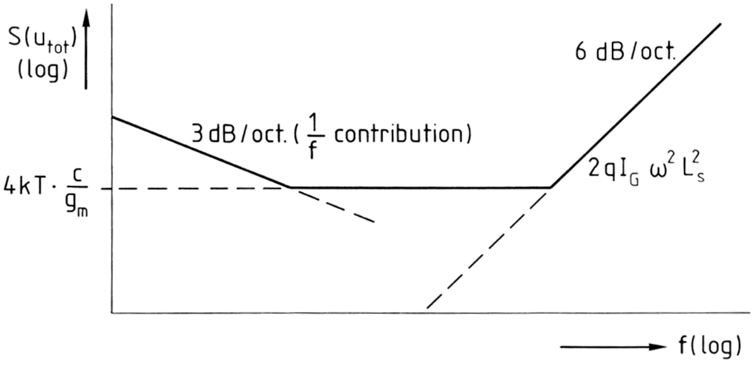


Figure 3.34. Spectrum of $S(u_{tot})$ (PET) with inductive source impedance. The 1/f contribution dominates in the low-frequency region.

From these figures it can be concluded that a bipolar transistor can be a better choice than an FET if the noise in the low-frequency region is of considerable interest.

3.9 Discussion

In this chapter we dealt with design considerations for the optimum noise performance of wide-band negative-feedback amplifiers. It was pointed out that the application of the proper type of negative feedback for impedance adaptation — avoiding large series impedances and small shunt impedances at the input — is a prerequisite for good noise performance.

Feedback with non-energic elements (gyrator and transformer) would be preferred, but can be approximated only in a few cases where transformers exclusively are used in the feedback network. The practical single-loop basic amplifier configurations with passive impedance-feedback networks were examined. Signal-to-noise ratio appears to be influenced as if the impedances, used to realise the feedback, are in parallel or in series with the input. It was shown that even in the case of lossless feedback the SNR may be deteriorated.

Subsequently, the influence on the noise performance of active feedback networks was investigated for some representative examples. It was pointed out that the active part of the network may have an adverse influence on the SNR when not properly designed.

By making use of the classification accession of the single-device configurations of chapter 2, criteria were formulated for the choice of the configuration of the input stage. Though the CE stage and the non-energic local-feedback stages (CB and CC stage) have nearly equal input noise sources, the CE stage must be preferred at the input, because its transfer parameters have the largest values, and as a consequence, a minimum noise contribution of the second stage is obtained. Other single-device configurations have larger input noise sources, and in addition at least one of their transfer parameters is reduced.

A design procedure was presented for realising optimum noise performance in wide-band amplifiers where no coupling transformers can be tolerated, so that the amplifier must be adapted to the source for maximum SNR and not vice versa. The procedure was worked out in some detail for various types of signal sources. Criteria were formulated for realising optimum unweighted signal-to-noise ratios.

The difference between the behaviour of a junction FIT and a bipolar transistor with respect to excess noise was emphasised. The FIT generally turns out to be most appropriate for capacitive sources. The bipolar transistor will frequently be the best choice for inductive sources.

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4 Design Considerations Regarding Optimum Accuracy and Linearity of Negative Feedback Amplifiers

4.1 Introduction

In chapter 3 we dealt with the noise performance of negative-feedback amplifiers and we concluded that it is largely determined by the properties of the first amplifier stage, provided that this stage and the feedback network are well designed. This chapter deals with some design aspects regarding the other amplifier stages.

Though bandwidth and stability considerations will not be discussed in this chapter, we should point out at this stage that they set an upper limit to the number of active devices in the signal path of an overall negative-feedback amplifier used between given signal source and load. For the purpose of finding design criteria with respect to accuracy and linearity we will presume that in every particular design situation the allowable number of stages is fixed, but that the stage configurations are free to be chosen with some restrictions as to the signs of their transfer functions.

The design objective is to realise maximum accuracy and linearity in this situation, by optimum use of the active devices. It will appear in chapter 5, which deals with the high-frequency performance of amplifier stages, that our starting point is correct only if the definition of a stage is adapted appropriately.

Both accuracy and linearity are quality aspects which have to do with variations of the parameters of the amplifier devices. We will ascribe inaccuracy to deviations of the parameters from typical values. Non-linearity is caused by signal-induced parameter variations.

Inaccuracy and non-linearity can partly be caused by variations in the parameters that determine the asymptotic gain $A_{t\infty}$. This is the case with series feedback with floating amplifier ports and with active-feedback amplifiers. Active-device parameters occur in these situations in the expressions for $A_{t\infty}$. These influences were disregarded until now, but in this chapter we will deal with them in some detail.

Because for the realisation of high accuracy largely the same criteria as for distortion hold, we will restrict ourselves mainly to linearity considerations. One important difference is that accuracy is determined by deviations of nominal parameter values in all stages, whereas distortion generally originates mainly in the output stage, because there the signal currents and voltages have values relatively large with respect to the bias currents and voltages. The other stages can usually be biased such that the relative current and voltage excursions are much smaller. Furthermore, the parameter variations due to the signal are fixed and correlated, whereas parameter variations due to various other causes cannot always be predicted and as a worst-case approach are regarded here as stochastic and uncorrelated. A brief discussion on accuracy will be given in section 4.6.

A profound treatment of distortion lies far beyond the scope of this work. An analysis of the distortion performance of non-linear circuits gets complicated if it encompasses the calculation of harmonics higher than the second, or distortion at high frequencies, even if only one amplifier stage is involved. We refer to various papers on this subject [1], [2], [3], [4].

We do not intend to develop an instrument for the prediction of quantitative distortion performance. Our aim is to find the configurations suited to use within a feedback loop for the attainment of optimum or near-optimum distortion performance. For this purpose the results of analytical work as described in the above-mentioned papers are not applicable, because they are not sufficiently general for the structures of the active parts that emerge from the design method developed in this work.

4.2 General considerations

It is useful to make a distinction between the coarse non-linearities caused by signals so large that certain active devices no longer operate in their ‘linear’ region and, on the other hand, smaller non-linearities caused by relatively small signal-induced excursions from the quiescent currents and voltages.

The first type of distortion must be avoided by making the bias currents and voltages sufficiently large. In the design of signal transfer this plays a role only in so far that the minimum values of currents and voltages, especially in the last stages of the amplifier, are determined on the basis of this criterion. To this type of distortion – besides low-frequency signal clipping – such specific phenomena as slewing and transient intermodulation distortion must be reckoned.

When bias currents and voltages in the amplifier are made sufficiently large for avoiding clipping, this may have consequences for the high-frequency performance. For operational amplifiers, where the ‘high-frequency’ response of the active part is usually determined by one dominant pole and a 20 dB/decade roll-off up to unity voltage gain, this relation was discussed by Solomon (slewing) [5], while Garde [6] dealt with transient intermodulation in some types of audio amplifiers. In the special-purpose amplifiers discussed in this work, there are many degrees of freedom for selecting adequate high-frequency compensation techniques to obtain the desired high-frequency response. Consequently, no generalised model can be given. We will pay some attention to the high-frequency compensation techniques in chapter 5. In chapter 7 a concise discussion of the subject will be given.

The second type of distortion which can be indicated as ‘small-signal distortion’ can generally be reduced by enlarging bias currents and voltages. Reducing distortion in this way, especially when pursued in the output stage, results in lower amplifier efficiency. For this reason, this type of quality improvement — which was resorted to in early days when feedback was unknown [7] — should be considered only when it is impossible to reduce distortion by selecting a better configuration, or by enlarging the loop gain.

‘Small-signal’ distortion performance of amplifiers is generally specified in terms of higher-harmonics generation for a sine-wave input signal with a given amplitude, or as intermodulation distortion for two input sine waves of different frequencies and given amplitudes. More recent methods of distortion specifications make use of second- and third-order intercept points to characterise intermodulation performance.

Seen from a designer’s point of view, harmonic-distortion specifications are difficult to handle as design objectives, because they depend in a complicated way on the non-linearities of the active devices and their environment. Moreover, these non-linearities are not known in suitable forms. We will therefore base our distortion (and accuracy) considerations on a different method.

4.3 Characterisation of non-linearity by means of the differential error

A convenient way of dealing with non-linearity for design purposes is given by Cherry and Hooper [8]. For defining the divergence of a transfer function from ideal linearity, they introduce the so-called *differential error:*

$$\varepsilon = \frac{TF - TF(Q)}{TF(Q)} \tag{4.1}$$

$TF(Q)$ is the magnitude of the transfer function in the quiescent point, while TF is the magnitude of this function for an actual signal value.

The transfer function can be characterised in terms of small-signal parameters, and for a prediction of the non-linearity the small-signal parameters must be known as functions of voltages and currents.

For a complete characterisation of non-linearity, ε must be known as a function of the signal excursion from the quiescent point. From such a relation, the more usual specifications can be deduced. The second- and third-harmonic distortion can be approximately calculated, however, from the difference and sum of ε^+ and ε^-, respectively, these being the values of ε for the peak positive and negative excursions from the quiescent point. If the 'small-signal' non-linearity has a smooth character, the magnitudes of higher harmonics rapidly decrease, and a specification of the lower harmonics will generally be sufficient.

In order to find design criteria for the stages in the feedback loop leading to optimum performance of the amplifier, we will first relate the differential error of the amplifier transfer function to that of the loop gain by writing:

$$\varepsilon_{A\beta} = \frac{A\beta - A\beta(Q)}{A\beta(Q)}$$

or

$$A\beta = A\beta(Q)(1 + \varepsilon_{A\beta})$$

where $\varepsilon_{A\beta}$ is defined, of course, for the actual excursions from the quiescent point at those terminals in the loop that correspond to the output terminals of the amplifier.

The differential error ε_{At} of the transfer function A_t can now be written as:

$$\varepsilon_{At} = \frac{\varepsilon_\beta}{1 - (1 + \varepsilon_{A\beta})A\beta(Q)} \tag{4.2}$$

For large values of $|A\beta(Q)|$ and $\varepsilon_{A\beta} \ll 1$, (4.2) can be approximated as:

$$\varepsilon_{At} = \frac{\varepsilon_{A\beta}}{-A\beta(Q)} \tag{4.3}$$

which represents the ratio of the differential error of the loop gain and the loop gain itself.

A characteristic and noticeable feature of *overall negative-feedback* amplifiers with a sufficient amount of loop gain is that the output level is approximately fixed for a given input signal, as it is determined by the asymptotic-gain network, and is therefore largely independent of the gain of individual stages. Loop gain and its differential error, however, strongly depend on the gain and configurations of individual stages.

A reasonable assumption, at least for the time being, is that distortion originates in the output stage, which means that $\varepsilon_{A\beta}$ is the differential error of the output-stage transfer function. We can draw the following important conclusion in that case from (4.3):

The ratio of the differential error and the contribution to the loop gain of a stage can serve as a figure of merit for its suitability as an output stage in an overall negative-feedback amplifier.

Furthermore, we can conclude from (4.3) that for low amplifier distortion the amplifier stages preceding the output stage should contribute as much to the loop gain as possible. Therefore, the best stages to use are CE stages because of their large transfer parameters. This desideratum is not constrained by the requirements of optimum noise performance, but it may conflict with the stability requirements to be discussed in chapter 5.

In section 4.4 we will investigate the differential-error-to-gain ratio (D-G ratio) for various single-device configurations. A restraint will be imposed on the load impedance. We assume that it is linear and that its value is known. Moreover, we will assume that the feedback network forms a negligible load for the output stage.

4.4 Differential-Error-to-Gain Ratio in Single-Active-Device Configurations and in the Current Mirror

4.4.1 Introduction

In order to compare the various single-device configurations with respect to their suitability for use as output stages in overall negative-feedback amplifiers, we will calculate the ratios of differential error and gain for a situation where the stages are loaded with a linear impedance Z_ℓ and driven from a linear impedance Z_s. The latter condition is in accordance with the previously formulated assumption that all distortion originates in the output stage. As a reminder we note that the output level of a negative-feedback amplifier depends neither on the structure of the active part of the feedback loop, nor on the gain of individual stages, provided that the loop gain is large. The output level is fixed by the input level and the asymptotic gain.

In the above-described situation which transfer function of the output stage is considered is immaterial. We arbitrarily take its voltage gain as indicated in figure 4.1. The stage is described with its transmission parameters, and the differential-error-to-gain ratio will be expressed in the differential errors of the transmission parameters.

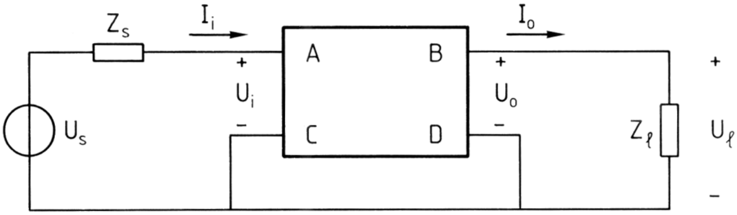

Figure 4.1. Twoport with source and load as a base for distortion consideration.

The voltage gain of this configuration is given by:

$$A_u = \frac{U_\ell}{U_s} = \frac{1}{A + B/Z_\ell + CZ_s + DZ_s/Z_\ell} \tag{4.4}$$

The differential error of the voltage gain is defined as:

$$\varepsilon(A_u) = \frac{A_u - A_{uQ}}{A_{uQ}} = \frac{\Delta A_u}{A_{uQ}} \tag{4.5}$$

The ratio of differential error (D) and gain (G) for small values of ΔA_u can therefore be written as:

$$\frac{D}{G} = \frac{\varepsilon(A_u)}{A_u} = \frac{\Delta A_u}{A_u^2} \approx -\Delta \frac{1}{A_u}$$

With the aid of expression (4.4) for the voltage gain A_u the D-G ratio can easily be written as a sum of terms:

$$\frac{D}{G} = \frac{\varepsilon(A_u)}{A_u} = -(\Delta A + \Delta B/Z_\ell + \Delta CZ_s + \Delta DZ_s/Z_\ell) \tag{4.6}$$

Note that expression (4.6) can be described in terms of the hybrid-*n* equivalent circuit if so desired. Numeric values can be found then, provided that the current and voltage dependencies of the parameters are known.

The variations in the transmission parameters with current and voltage are correlated, and they may have opposite signs so that compensation effects may occur for specific combinations of Z_s and Z_ℓ.

However, it is not considered wise to rely on a distortion cancellation due to compensation in a single-device configuration because of the following reasons. Unless the device parameters and their current and voltage dependencies are measured individually, they are not known sufficiently well (with the exception of the low-frequency transconductance of a bipolar transistor). Moreover, the compensation is very likely to be temperature dependent. Furthermore, if a cancellation is realised for, for example, even harmonics, it is not very likely that odd harmonics would be reduced as well, and inaccuracy would certainly not be reduced.

Therefore, in order to find operative design criteria, it makes sense to assume as a worst-case approach that the variations of the parameters in (4.6) all have equal signs and that no compensation effects will occur.

With this assumption we will first examine the influence of series and shunt impedances in the signal path on the D-G ratio. Subsequently, we will compare the D-G ratio of various single-device configurations with that of the CE stage. Only bipolar-transistor stages will be discussed, but the results are general and can be applied to corresponding FIT stages and to balanced configurations as well.

A comparison of FIT and bipolar-transistor stages would require detailed knowledge about the non-linearities of all parameters of both devices, a task beyond the scope of this work.

4.4.2 The influence of series and shunt impedances at the output on the D-G ratio

In chapter 1, we postulated as an initial design consideration that the insertion of series or shunt impedances at the output of an amplifier in order to realise the desired source impedance for the load is pernicious from the view point of power efficiency and distortion. We will investigate in this section the influence of such impedances on distortion in order to further support the criteria, as given in chapter 1, for the selection of the proper type of feedback at the output of an amplifier.

(i) Influence of a shunt impedance at the output

Figure 4.2 shows a twoport (CE stage) in connection with an output *shunt impedance*. The parameters E and D remain unchanged with respect to those of the CE stage, whereas the parameters A' and C' of the twoport within the dashed framework have the values:

$$A' = A + B/Z_o \qquad\qquad C' = C + D/Z_o$$

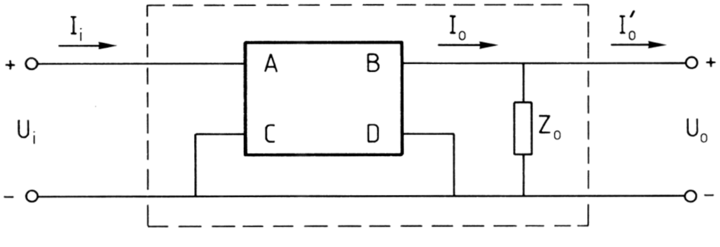

Figure 4.2. Twoport with shunt impedance at the output.

With these parameter values the expression for the D-G ratio of the voltage gain becomes:

$$\frac{D'}{G'} = \Delta A + \Delta B\left(\frac{1}{Z_o} + \frac{1}{Z_\ell}\right) + \Delta C Z_s + \Delta D Z_s\left(\frac{1}{Z_o} + \frac{1}{Z_\ell}\right) \tag{4.7}$$

From this expression it can be concluded that the contribution of ΔB and ΔD are enlarged unless parallel-resonant circuits are formed in such a way that $(1/Z_o+1/Z_\ell) < 1/Z_\ell$. Such a parallel resonance can reduce the D-G ratio in a restricted frequency region only and has no significance in wide-band amplifiers. Moreover, we can conclude that in order to maintain the same output level the shunt impedance necessitates larger current excursions in the active device, and consequently, the differential errors of the current-dependent parameters will be enlarged when the bias current is maintained at the same level. When the bias current is enlarged, for the purpose of reducing the relative current excursions, power efficiency decreases.

(ii) Influence of a series impedance at the output

For the case where an impedance is connected *in series* with the output as shown in figure 4.3, the parameters A and C remain unchanged, while B' and D' have values:

$$B' = B + AZ_o \qquad D' = D + CZ_o \tag{4.8}$$

The D-G ratio of the voltage gain in this case is therefore given by:

$$\frac{D'}{G'} = \Delta A\left(1 + \frac{Z_o}{Z_\ell}\right) + \Delta B/Z_\ell + \Delta C Z_s\left(1 + \frac{Z_o}{Z_\ell}\right) + \Delta D\frac{Z_s}{Z_\ell} \tag{4.9}$$

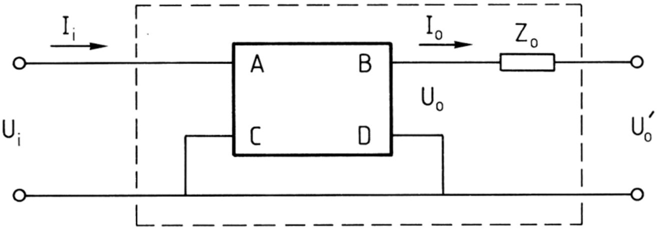

Figure 4.3. Twoport with series impedance at the output.

An increase of the contributions with ΔA and ΔC is observed. Moreover, the variation in the voltage-dependent parameters is enlarged because of the voltage drop across Z_o. When the bias voltage in the stage is increased so that the same relative voltage excursions are obtained the distortion may be reduced, but the power efficiency decreases.

On these grounds we conclude that the insertion of series or shunt impedances at the output of wide-band amplifiers must be rejected as a means to enlarge or reduce the output impedance of an amplifier.

4.4.3. The D-G ratio of stages with series and shunt impedances at the input

The insertion of series and shunt impedances at the input of an output stage might be considered for changing the character of the source impedance in order to obtain a more linear transfer of that stage. Though the transfer may be linearised by more pronounced voltage or current driving, this way of linearising results in an increased D-G ratio, as can be seen from expressions (4.10) and (4.11), which represent the D-G ratio in the case of a *series* impedance and a *shunt* impedance, respectively, both denoted by Z_i.

$$\frac{D'}{G'} = \Delta A + \Delta B/Z_\ell + \Delta C(Z_s + Z_i) + \Delta D(Z_s + Z_i)/Z_\ell \tag{4.10}$$

$$\frac{D'}{G'} = \Delta A\left(1 + \frac{Z_s}{Z_i}\right) + \frac{\Delta B}{Z}\left(1 + \frac{Z_s}{Z_i}\right) + \Delta C Z_s + \Delta D Z_s/Z_\ell \tag{4.111}$$

In the first case, the influence of the parameters C and D increases, but no reduction in the influence of the parameters A and B is found.

The insertion of a series resistance in the case of a CE output stage, for example, may linearise the transfer function because it is in a larger measure current driven. From expression (4.10) it follows, however, that when this linearisation technique is used in an output stage of a negative-feedback amplifier, *the advantage of a more linear transfer is wiped out by the reduction in its gain contribution*.

By inserting a shunt impedance, the influence of the non-linearities of B and A is enlarged, while that of C and D remains unchanged.

An alternative method for changing the drive impedance of the output stage is to apply local negative feedback to the driving stage. It can easily be shown for the combination of stages that the D-G ratio is always enlarged by doing so.

4.4.4 The influence of local negative feedback on the D-G ratio

Next, we will examine the influence of local negative feedback on the D-G ratio. The twoport of figure 4.1 is again supposed to be a CE stage, and expression (4.6) is used as a reference for comparison.

(i) D-G ratio of non-energic local-feedback stages

The application of non-energic feedback is expected to yield the best results, because series or shunt impedances are not needed. Figure 4.4 depicts the CC and CE stages with the twoport symbols of figure 4.1. For their parameters (indicated by primes and expressed in those of the CE stage), the following values are found:

CC stage:

$$A' = \left(1 - A - \frac{BC}{1-D}\right)$$

$$B' = -B - \frac{BD}{1-D}$$

$$C' = -\frac{C}{1-D}$$

$$D' = -\frac{D}{1-D}$$

CE stage:

$$A' = -\frac{A}{1-A}$$

$$B' = -\frac{B}{1-A}$$

$$C' = -C - \frac{AC}{1-A}$$

$$D' = \left(1 - D - \frac{BC}{1-A}\right)$$

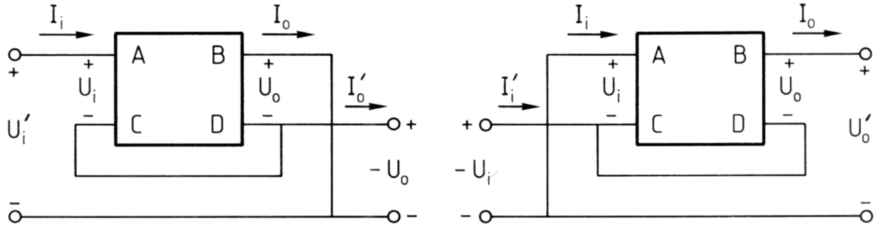

Figure 4.4 CC stage and CB stage represented by twoport symbols.

In chapter 2 we gave the approximate values of the transfer parameters and showed that for low frequencies only one parameter is changed significantly if single-loop feedback is used. This is confirmed by the parameter values given here.

The parameters A and D are small compared to unity over quite a large frequency range and can be disregarded where present in the denominators of the expressions. The parameter product BC is very small at low frequencies (about $1/(\alpha_{eo}\mu_{eo})$ and can be approximated by $BC \approx pC_{b'c}/g_e$ at higher frequencies. It is therefore small compared to D, which can be approximated by $D \approx -pC_{b'e}/g_e$ at high frequencies and is roughly equal to A, which has a high frequency value $A \approx -pC_{b'c}/g_e$. Consequently, D-G ratio of the CC stage and CE stage at relatively low frequencies equal that of the CE stage.

The differential error of the voltage-gain factor of the CC stage decreases at higher frequencies because of a cancellation effect due to the approximate equality of $-A$ and BC. The D-G ratio at high frequencies can therefore be somewhat better than that of the CE stage.

(ii) D-G ratio of impedance-feedback stages

Impedances used in the feedback networks of the series and shunt stages have effects on the distortion performance similar to the series and shunt impedances of figures 4.2 and 4.3. The expressions found for the D-G ratio are complicated, and do not lead to clear insight. Approximate expressions for the transmission parameters are given below:

series stage:

$A' = A + CZ$

$E' = Z(-1 + A + B/Z + D + BC - AD)$

$C' = C$

$D' = D + CZ$

shunt stage:

$A' = A + B/Z$

$B' = B + DZ$

$C' = \frac{1}{Z}\,(-1 + CZ + D + Z + BC - AD)$

$D' = D + B/Z$

The impedances Z in these expressions are the feedback-network elements. The parameters of the series stage have a common denominator $(1 + CZ)$. This denominator is approximated as unity, assuming $|CZ| \ll 1$, which corresponds to $Z \ll \alpha_{eo}\mu_{eo}r_e$ at low frequencies and to $|Z| \ll 1/\omega C_{b'e}$ at higher frequencies. A similar approximation is used with the shunt stage, where it is assumed that $|B/Z| \ll 1$ $(g_e Z \gg 1)$.

When for finding the low-frequency D-G ratio of these configurations, the terms AD-BC in the parameters are neglected with respect to unity, we find the following expressions for the series and the shunt stages, respectively.

$$\frac{D'}{G'} = \Delta A\left(1 + \frac{Z}{Z_\ell}\right) + \Delta B\frac{1}{Z_\ell} + \Delta C\left\{Z_s + Z\left(1 + \frac{Z}{Z_\ell}\right)\right\} + \Delta D\frac{(Z + Z_s)}{Z_\ell} \qquad (4.12)$$

$$\frac{D'}{G'} = \Delta A\left(1 + \frac{Z_s}{Z_\ell}\right) + \Delta B\left\{\frac{1}{Z_\ell} + \frac{1}{Z}\left(1 + \frac{Z_s}{Z_\ell}\right)\right\} + \Delta C Z_s + \Delta D\left(\frac{Z_s}{Z} + \frac{Z_s}{Z_\ell}\right) \qquad (4.13)$$

The D-G ratios of the parameters stabilised by the feedback action appear to be unchanged, but the other contributions to the total D-G ratio are enlarged. This will generally result in a deterioration in accuracy and distortion performance.

An additional increase in the D-G ratio will be brought about by the enlarged signal that must be handled by the active device as a consequence of the output power loss in the feedback network in order to produce the same output signal as in the case of a CB stage. This power loss can, of course, be much smaller than in the case where, for obtaining the same effect on the output impedance, a series or shunt impedance at the output of a CE stage would be used.

4.4.5 D-G ratio of indirect-feedback output stages

As an example of an indirect-feedback output stage, we will examine the performance of a current mirror with bipolar transistors, schematically shown in figure 4.5.

Even when the devices are exactly equal, the parameters cannot be assumed to be so, because output currents and voltages of both devices may differ. The parameters of the complete twoport have the values:

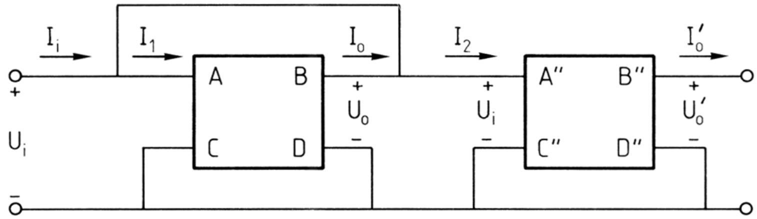

Figure 4.5. Twoport representation of a current mirror.

$$A' = A''$$

$$B' = B''$$

$$C' = C'' + CA'' + A''(D-1)\,(1-A)/B$$

$$D' = D'' + CB'' + B''(D-1)\,(1-A)/B$$

Because B'' and B are very nearly equal in the case of equal bias currents (equal transconductances) the parameter D' has a value close to unity. Ignoring small contributions, the D-G ratio has a value

$$\frac{D}{G} = \frac{\varepsilon(a)}{A_u}$$

$$= \Delta A'' + \frac{\Delta B''}{Z_\ell} + \Delta C'' Z_s + \frac{\Delta D'' Z_s}{Z_\ell} + (\Delta A + \Delta D)\frac{Z_s}{Z_\ell} + \frac{Z_s}{B}\left(\Delta B \frac{A''}{B} - \Delta A''\right) \quad (4.14)$$

which may be considerably larger than that of a CE stage.

4.4.6 Some concluding remarks

Surveying the results of the foregoing subsections, we can conclude that the D-G ratio of a stage – being a figure of merit for a configuration to determine its suitability as an output stage in an overall negative-feedback amplifier – is deteriorated by inserting series or shunt impedances at its input or output. The D-G ratio of local-feedback stages will be equal to or larger than that of the CE stage as well.

An additional phenomenon of applying negative feedback to the output stage is that the driving stage has to handle larger signals to maintain the same output level because one or more transfer parameters are reduced. Consequently, the assumption of its linearity may no longer be correct. As a result, the differential error of the amplifier transfer is increased in all probability unless parameter variations in the output stage are accurately compensated by opposite variations in the driving stage. Such compensations are usually believed to be unreliable unless they refer to one and the same parameter, the magnitude of which is varied by exactly the same amount. These types of compensation will be concisely dealt with in the next section.

4.5 Compensation techniques

It is not so obvious that the use of local feedback in the output stage should always be avoided for minimising distortion. The reduction in an output stage transfer parameter by means of local feedback may result in a distortion reduction, not because of an improved D-G ratio of the stage itself, but because a compensation caused by the distortion in the corresponding parameter of the driving stage may occur. This can be seen from a transmission parameter representation of the cascaded driving and output stages:

$$\begin{aligned} A &= A_1A_2 + B_1C_2 \qquad & B &= A_1B_2 + B_1D_2 \\ C &= C_1A_2 + D_1C_2 \qquad & D &= C_1B_2 + D_1D_2 \end{aligned} \tag{4.15}$$

The subscripts 1 refer to the driving stage and the subscripts 2 to the output stage.

A reliable even-order compensation is obtained when a transmission parameter of the cascaded combination is determined either by the product or by the sum of two corresponding parameters, provided that these parameters and their differential errors have equal magnitudes, but the latter have opposite signs.

An example, illustrating the compensation mechanism, is the cascade of a CC stage and a CE stage, where a linearisation of the transconductance $1/B$ is obtained because of the unity values of the parameters A_1 and D_2, respectively. The bias currents in both stages have to be equal and their sum has to be constant. Other cascaded combinations exhibiting similar compensations can be found on the basis of expression (4.15).

4.6 Some accuracy aspects

The influence of parameter variations on amplifier accuracy can be described in a way similar to that done above for distortion (Eq. (4.1)) in section 4.3. In that case $TF(Q)$ must be interpreted as the magnitude of the transfer function for certain typical parameter values at a specified temperature and time.

The differential error is the relative deviation from this value caused by parameter spread, ageing, temperature changes and changes of supply voltage, etc. As a worst-case approach, it can be assumed that the parameter variations are stochastic and uncorrelated. Unlike non-linearities, inaccuracy is determined by parameter variations in *all* amplifier stages. Therefore, we can conclude that for optimum accuracy the D-G ratios of all stages should be minimised. This requires the avoidance of series as well as shunt impedances and of local feedback.

The realisation of higher accuracy implies that input and output impedances better approximate the ideal zero or infinite values in single-loop amplifiers. It may be clarifying to point here at a possible line of incorrect reasoning that tempts a designer into trying to realise a small output impedance in a negative-feedback amplifier by using a shunt impedance or local shunt feedback at the output of the output stage. This technique appears to have little effect on the output impedance, but instead causes the input impedance to deviate more from its intended value as this shunt resistance gets smaller or the amount of local feedback larger, because the loop gain is reduced. Similar reasoning applies for similar measures at the input in order to increase or reduce the input impedance.

4.7 Non-linearity and inaccuracy of the asymptotic gain in the case of floating amplifier ports (series feedback)

4.7.1 Introduction

The expressions used until now for the asymptotic gain of all amplifier configurations were based on the assumption that the active parts behave as perfect nullors. We mentioned in chapter 1, section 1.4.4 that the imperfect isolation from ground of the input and output ports of the active part leads to departures from the ideal asymptotic-gain value. Consequently, the asymptotic gain is not entirely determined by the components of the feedback network.

Following our program as outlined in section 4.1, we will now analyse these situations, which can be accomplished to advantage again with the aid of transmission-parameter representations of the input and output configurations.

4.7.2 Inaccuracy due to series feedback at the input

First we will analyse the influence of an imperfectly floating input port in the case of input series feedback. The input configuration may comprise one or more active devices. The complete configuration is assumed to have one output terminal in common with an input terminal. This situation is depicted in figure 4.6, where the input configuration is indicated separately in a voltage amplifier.

Formally, the asymptotic gain $A_{t\infty}$ can be determined by applying the nullor constraints to the terminals i and i' of the input stage, i.e. by supposing U_i and I_i to be zero. The calculation, however, is complicated by the fact that I' does not become zero, and therefore, the transfer function T enters in the expression for $A_{t\infty}$. Because T is an unspecified part of the amplifier this is undesirable. Our aim is to find design criteria for the first stage.

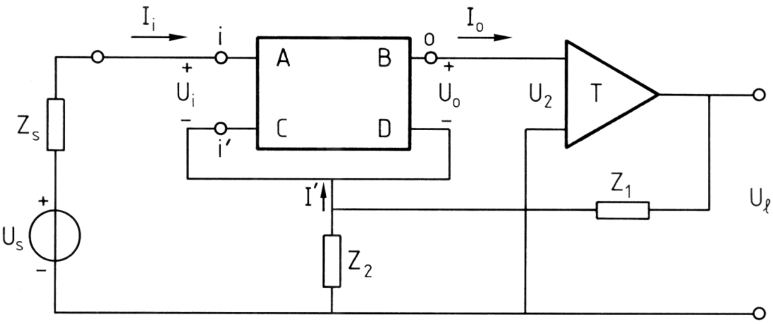

Figure 4.6. *First stage indicated separately in a voltage-amplifier configuration.*

The nullor constraints are preferably applied to the input of the symbolic amplifier T, while U'_ℓ and U_s are taken as the input voltages of the input stage, as indicated in figure 4.7. It follows that:

$$A_{t\infty} = \left(\frac{U_\ell}{U_s}\right)_{T\to\infty} = \frac{Z_1 + Z_2}{Z_2}\;\frac{1 + CZ'}{1 - A - CZ_s}$$

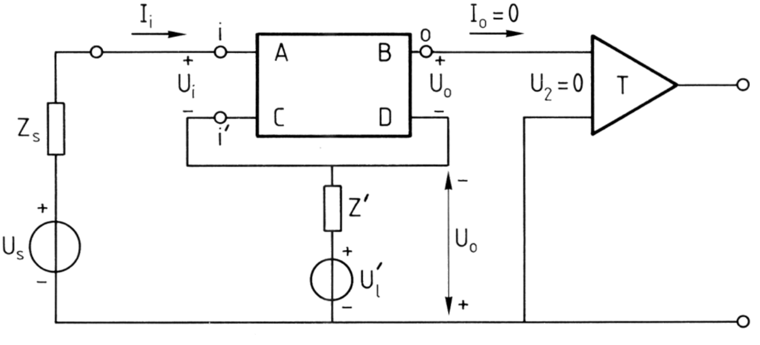

Figure 4.7. *Equivalent circuit for the determination of* A_t
$U'_\ell = U_\ell Z_2/(Z_1 + Z_2)$, $Z' = Z_1Z_2/(Z_1 + Z_2)$.

Only the parameters A and C of the input device occur in the expression for $A_{t\infty}$ now. They represent the effects of an imperfect isolation of the input port from ground and should have values as close to zero as possible. It may be clarifying to examine their influence for the familiar case of a single transistor input configuration.

For a circuit with a single transistor input stage, where the base corresponds to terminal i, the emitter to i' and the collector to 0, the parameters A and C are found to have the values:

$$A = -\frac{1 + p\mu_{eo} r_e C_{b'c}}{\mu_{eo}} \qquad C = -\frac{1 + p\alpha_{eo}\mu_{eo} r_e C_{b'c}}{\alpha_{eo}\mu_{eo} r_e}$$

when the influence of singularities far from the origin is disregarded.

The finite values of μ_{eo} and $1/pC_{b'c}$ thus limit the accuracy of the asymptotic gain even when the loop gain with respect to T becomes infinite! For bipolar transistors μ_{eo} has a value between 10^3 and 2×10^4. This means that – at low frequencies – only in amplifiers that must be extremely accurate and linear does its finite value play a role.

When an FET is used as an input device, the situation is less favourable. The quantity $\mu_{so} = g_m r_{ds}$ then has a value between 50 and 200. Especially with a low drain-to-source voltage, the circuit has a bad accuracy and the distortion can be large.

At higher frequencies and in the case of a capacitive signal source, $C_{b'c}$ influences accuracy and linearity. This can be seen immediately from figure 4.7: if $T \to \infty$, U_2 becomes zero and $C_{b'c}$ is virtually shunting the input of the amplifier.

The effects of μ_{eo} and $C_{b'c}$ can be reduced to a large extent, at least at low frequencies, by the application of a 'bootstrapping' technique. An example is given in figure 4.8, where the input stage is a cascode. It can easily be shown that the parameter C is effectively reduced in this case, as is the lowfrequency value of A. At high frequencies this parameter is virtually unaltered.

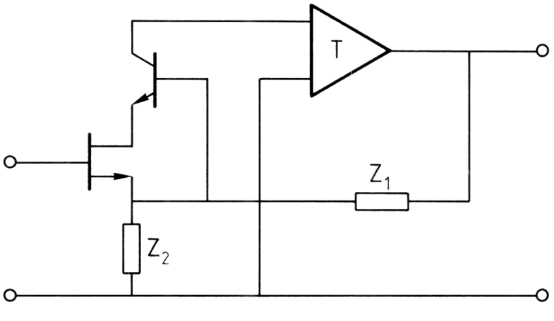

Figure 4.8. Cascode input stage to reduce the effect of non-zero values of the parameters A and C of the input device.

Noise performance is not significantly affected by using a cascode instead of a single transistor. In chapter 5 it will be shown that the influence of the extra active device in the input stage usually does not have an adverse effect on bandwidth and stability either.

Application of a differential pair at the input yields an accurate alternative. Here, the commonmode-rejectionratio (CMRR) sets a limit to the accuracy. A disadvantage of a differential-pair input stage is that it generally has an adverse effect on noise performance.

4.7.3 Inaccuracy due to series feedback at the output

The application of series feedback at the output, with a single transistor in the output stage, also leads to a limited accuracy because of imperfect isolation of the output port to ground.

Figure 4.9 shows as an example a current-amplifier configuration where the output configuration is described again with its transmission parameters. The nullor constraints can be applied to the input of T, i.e. U and I are assumed to be zero. Then the analysis is reduced to the determination of the transfer function $(I_\ell/I_s)_{T\to\infty}$ for the circuit given in figure 4.10. It follows that:

$$\left(\frac{I_\ell}{I_s}\right)_{T\to\infty} = \frac{Z_1+Z_2}{Z_2}\,\frac{1+CZ'}{1-D-CZ_\ell}$$

The non-zero values of the parameters C and D limit the accuracy of the asymptotic gain in this case.

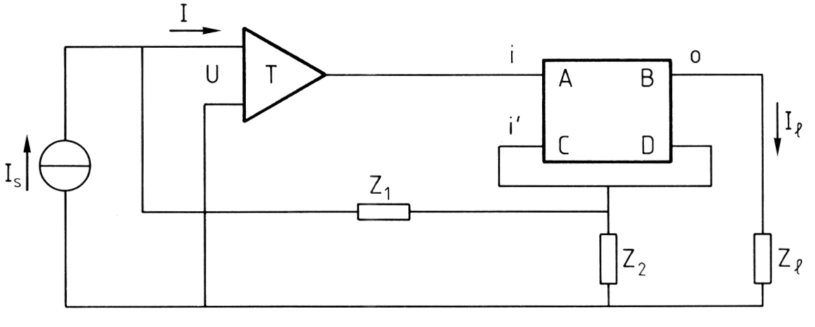

Figure 4.9. Equivalent circuit of current amplifier; the output stage is indicated separately.

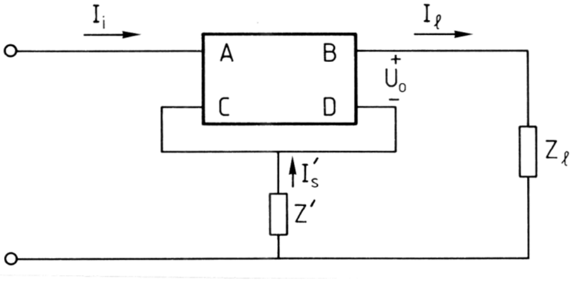

Figure 4.10. Equivalent circuit for the determination of $A_{t\infty} = (I_\ell/I_s)$, $Z' = Z_1Z_2/(Z_1 + Z_2)$, $I'_\ell = I_s(Z_1 + Z_2)/Z_2$.

We will examine again the case of a single-transistor output stage, where — referring to figure 4.9 — the emitter corresponds to terminal i', the base to i and the collector to 0. Collector and emitter may be interchanged without altering the results significantly (because the parameters C and D do not differ significantly).
The parameters C and D are found to have values

$$C = -\frac{1 + p\alpha_{eo}\mu_{eo}r_e C_{b'c}}{\alpha_{eo}\mu_{eo}r_e} \qquad D = \frac{1}{\alpha_{eo}} + \frac{j\omega}{\omega_T}$$

The finite value of the transistor parameter α_{eo} limits the low-frequency accuracy, whereas the finite value of ω_T and the non-zero value of $C_{b'c}$ limit the accuracy at high frequencies. Low-frequency inaccuracy can obviously be reduced by taking an FET output stage instead of a bipolar transistor. Highfrequency accuracy can usually not be improved in this way because the ω_T of an FET is usually not larger, nor will C_{dg} be smaller than $C_{b'c}$.
A better limited accuracy can be obtained here as well by using a differential-pair output stage as depicted in figure 4.11. This situation is not adequately described by the diagram of figure 4.9, because there is no interconnection between an input and output terminal. We will not analyse this situation in detail. It can easily be seen from figure 4.11, however, that at low frequencies the output current and the feedback current can differ much less than with a single output transistor as a result of base-current compensation.

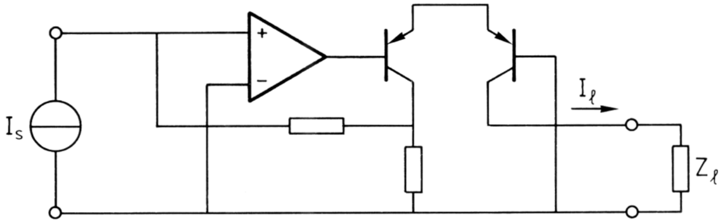

Figure 4.11. Current amplifier with differential output stage.

4.7.4 Concluding remarks

In this section we have considered the imperfections in the active part of negative-feedback amplifiers in as far as they influence the expressions for the asymptotic gain as a consequence of an imperfect isolation of the input or output ports.
The considerations with regard to accuracy and distortion are relevant only in cases where very large loop gains must and can be realised or where very large source and load impedances are involved (because of the shunting effect of $C_{b'c}$ at the input as well as at the output).
Of course, it is not possible to enlarge loop gain without limit. Very often one has

to be satisfied with a moderate loop-gain value, because singularities of the loop transfer function do not allow large values.

The methods discussed for enlarging the limiting accuracy in the case of series feedback are only useful for a restricted number of applications. In many cases, the assumptions $\mu_{eo} \rightarrow \infty$ and $C_{b'c} = 0$ lead to a sufficiently accurate approximation of the asymptotic gain.

4.8 Linearity and accuracy of amplifiers with active feedback

4.8.1 Introduction

In chapter 1 we distinguished two fundamentally different methods for realising approximately linear transfer functions with the aid of active feedback networks. The first is referred to as an indirect-feedback technique and uses dummy input and/or output stages. The second, called an active-feedback technique, uses direct sensing and comparison at output and input respectively.

The active feedback networks have the obvious disadvantage that they introduce non-linearities in the asymptotic gain. We will briefly examine in this section the performance of these networks in order to provide an impression of their practical usefulness.

4.8.2 Linearity and accuracy considerations for indirect-feedback techniques

In chapter 1 we dealt with indirect-feedback amplifiers with ideal input and output stages. In the amplifier with indirect current sensing at the output, we assumed in most cases that the voltage-gain factors of the output stages were sufficiently large, so that they did not significantly influence the transfer functions. In the amplifier with indirect voltage comparison we made a similar assumption for the current-gain factors of the input stages.

In this section we will more closely examine an indirect-feedback current amplifier and a voltage amplifier where the transfer parameters of the stages are all finite, and try to find some measures for improving linearity.

(i) Indirect-feedback current amplifier

Figure 4.12 shows the indirect-feedback current amplifier. The asymptotic value of the current gain is:

$$A_{i\infty} = \frac{I_\ell}{I_s} = -\left(1 + \frac{Z_2}{Z_1}\right)\left(\frac{AZ' + B}{A'Z_\ell + B'}\right) \qquad (4.16)$$

where $Z' = Z_1Z_2/(Z_1 + Z_2)$.

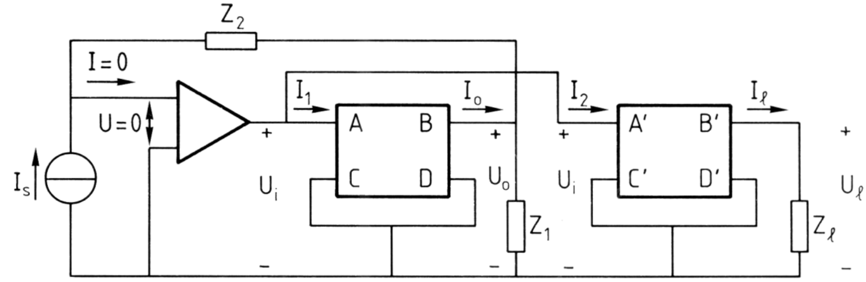

Figure 4.12. Indirect-feedback current amplifier.

When the bias currents in both stages are equal, an almost perfect linearity can be obtained if the additional condition that $Z' = Z_\ell$ is met. For Z_2 large with respect to Z_1, this means that Z_ℓ and Z_1 must be equal. In that case U_o and U_ℓ are equal as well, and as a consequence A' and A can be equal. A disadvantage of meeting this condition may be, of course, that both output stages must handle equal signal powers. Moreover, Z_ℓ must be known relatively well. Especially the latter requirement may be a serious drawback.

As we have mentioned already in chapter 1, the non-linearity of the transfer can alternatively be ascribed to the non-linearity of the output impedance of the output stage ($Z_o = B/A$ under voltage-drive conditions). An enlargement of this impedance leads to an improvement in the linearity.

There are two methods for realising a larger output impedance under voltagedrive conditions. When we take a CE stage as a starting point, we can either add a second stage or apply negative feedback.

The first method can be implemented by adding a stage that doesn't change the parameters B but that reduces the parameter A. A CE stage satisfies this requirement, and consequently, cascode output stages will perform better than CE stages.

The second method requires the application of the proper type of negative feedback increasing the parameter B and maintaining the value of A. The series stage described in chapter 2 satisfies these conditions and is therefore a simpler alternative.

In the indirect-feedback current amplifier without passive feedback components, the expression for A. becomes:

$$A_{i\infty} = \frac{I_\ell}{I_s} = -\frac{B}{A'Z_\ell + B'}$$

To reduce the influence of the parameter A' one of the above methods can be used.

(ii) Indirect-feedback variable-gain current amplifier

In chapter 1 we mentioned the possibility of realising a variable-gain amplifier by using differently biased bipolar-transistor differential pairs as output stages (figure 2.24). This possibility arises from the fact that the transconductance of a bipolar transistor is linearly proportional to the bias current, at least at low frequencies.

In chapter 2 we found for the transadmittance of a CE stage:

$$\gamma = \frac{g_b(Y_f - g)}{Y_i + Y_f + g_b} \approx \frac{-1}{r_e + r_b\left(\frac{1}{\alpha_{eo}} + \frac{j\omega}{\omega_T}\right)}$$

disregarding the right-half-plane zero.

It can be seen from this expression that the non-zero value of r_b may adversely influence linearity at low frequencies while at high frequencies the finite transit frequency deteriorates the linearity even more seriously.

The finite values of the parameters A and A' also play a role in these amplifier types. They can be reduced, if so desired, by using cascoded output stages. Negative feedback can, of course, not be used in this case because the required proportionality of the transconductance with the bias current gets lost.

(iii) Indirect-feedback voltage amplifier

An indirect-feedback voltage amplifier is depicted in figure 4.13. The active devices of the input stages are assumed to be well matched.

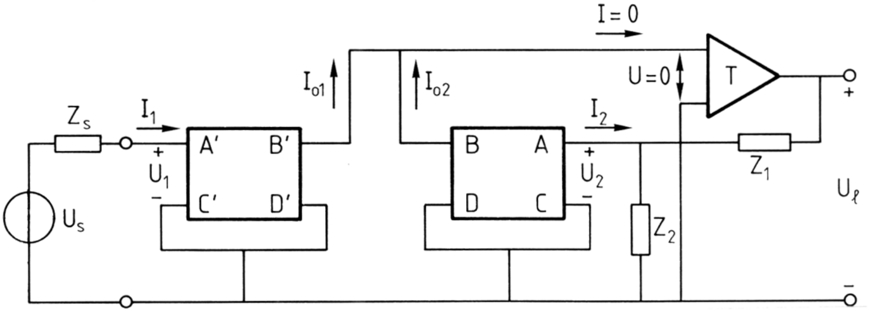

Figure 4.13. Indirect-feedback voltage amplifier.

The asymptotic value of the voltage gain ($T \to \infty$) is given by:

$$\frac{U_\ell}{U_s} = -\left(\frac{Z_1 + Z_2}{Z_1}\right)\frac{B + DZ'}{B' + D'Z_s}$$

where $Z' = Z_1Z_2/(Z_1 + Z_2)$.

Provided that the bias currents are equal, perfect linearity can be obtained when $Z_s = Z'$ and the parameters vary in the same way along with the output currents, so that for all signal values $B = B'$ and $D = D'$. Because the output currents are opposite, the parameter deviations must either be equal for both negative and positive signal excursions (which requires a symmetrical transfer characteristic), or the deviations must have opposite signs (which requires complementary transfer characteristics). We gave some examples of simple input stages having these properties in chapter 1. A disadvantage of meeting the condition $Z_s = Z'$ is, of course, that a relatively large amount of noise may be added. Moreover, Z_s must be well known and linear.

A more preferable way of obtaining linearity is based on reducing the influence of the parameter D, the reciprocal current-gain factor. Taking the CE stage as a reference (figure 4.14) two techniques are available.

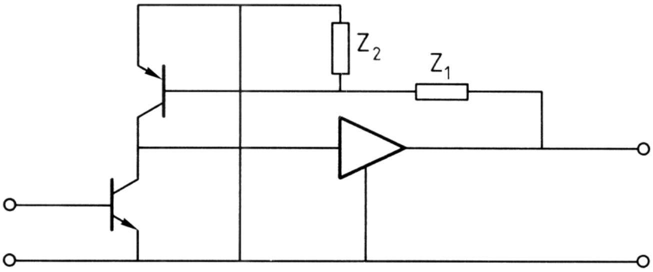

Figure 4.14. Indirect-feedback voltage amplifier with complementary CE input stages.

The first technique requires the use of a device with an inherently larger current-gain factor such as an FET as depicted in figure 4.15, or a combination of stages with an enhanced current-gain factor and a virtually unaltered transconductance, such a Darlington pair.

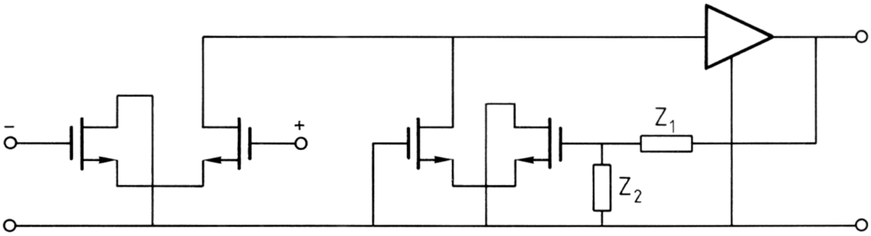

Figure 4.15. Indirect-feedback voltage amplifier with symmetrical FET input stages.

The second technique uses negative feedback which maintains the current gain at the same value, while reducing the transconductance. The series stage (either balanced or unbalanced) satisfied these requirements. To put it another way, its input impedance with a short-circuited output is larger than that of the CE stage.

In the indirect-feedback voltage amplifier, where gain scaling is achieved by scaling the number of series-connected devices in stage II of figure 4.16, the expression for $A_{u\infty}$ becomes:

$$A_{u\infty} = -n\left(\frac{B'}{B + DZ_s}\right)$$

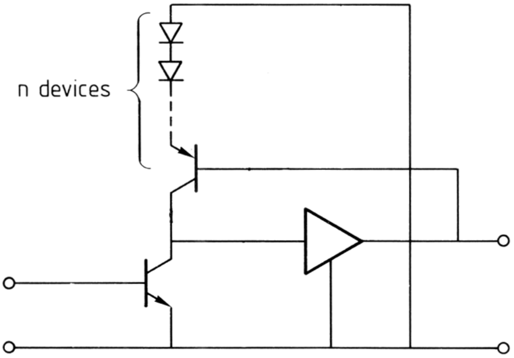

Figure 4.16. Indirect-feedback voltage amplifier with gain scaling achieved by using several series-connected devices in stage II.

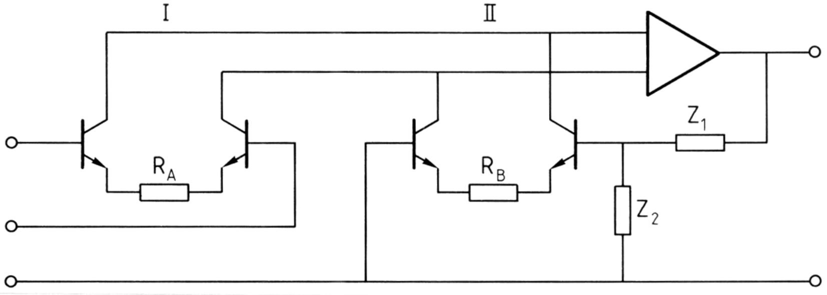

Figure 4.17. Indirect-feedback voltage amplifier where the voltage gain is determined by the ratio of Z_1 and Z_2 as well as by that of R_B and R_A.

Either of the above techniques can be used to reduce the influence of the parameter D.

One might try to achieve a gain scaling by giving the input stages different transconductances by providing them with unequal degeneration resistances R_A and R_B, as depicted in figure 4.17. Perfect linearity cannot he obtained in this way unless in stage II a number of devices equal to the ratio of R_B and R_A is connected in series and bias currents in all stages are equal.

The linearity can of course be improved alternatively by using more elaborate input stages. Negative feedback, compensation, prior-feedforward techniques or combinations of these may he considered for improving the linearity of the *individual* transconductances. An elaboration of such techniques is beyond the scope of this work.

4.8.3 Linearity and accuracy considerations for active-feedback techniques

In chapter 1 we presented as examples of active-feedback techniques some amplifiers that used inverting current attenuators to set the current-gain factor. Three possibilities were mentioned for the implementation of these attenuators. We will in this section discuss some design aspects in relation to the non-linearity of these networks. For other amplifier configurations using active-feedback techniques similar considerations apply.

Because the active feedback networks have to function as attenuators, the invertors used in them can operate under rather ideal drive and load conditions. The loop gain in the feedback configurations can therefore he rather large even in the case where local-feedback configurations are used.

The three alternatives of chapter 1, section 1.5.3 are depicted in figure 4.18 with these local-feedback configurations in their feedback networks.

Figure 18a uses a series stage in the feedback network driven by a voltage proportional to the amplifier output current. In order to keep the non-linearity of the series-stage transfer small, the voltage across R_1 must be small. The resistor R_2 is selected so that the required value of α is realised ($\alpha = -R_2/R_1$), and consequently, R_2 attains a relatively small value. This, of course, adversely influences noise performance.

The second configuration (figure 4.18b) uses a shunt stage in its feedback network. This stage has to handle the complete output current. When the current is large, this may cause considerable distortion. A larger value of R_1 causes the output voltage of the stage to increase. This may cause the distortion to increase as well. The resistor R_2 can be made large to obtain good noise performance. Configuration 4.18b is more favourable in terms of noiseperformance. Configuration 4.18a will perform better as far as linearity is concerned.

This means that configuration 4.18a is preferably used in combination with a direct-feedback loop in order to realise an accurate and linear output impedance. An accurate low-noise input impedance can better be realised with configuration 4.18b extended with a second direct-feedback loop.

The configuration of figure 4.18c is suitable for realising an accurate output impedance. Its noise performance is not very attractive. Applications lie in the domain of integrated-circuit cable-drive amplifiers.

In general, it can be concluded in qualitative terms that the ‘closer’ the active device (or a more complicated active part) of the active feedback network is to the output, the larger the non-linearity will be. Noise performance deteriorates the closer the active device gets to the input. This general, but perhaps rather trivial, rule applies equally well to other configurations that use active-feedback techniques.

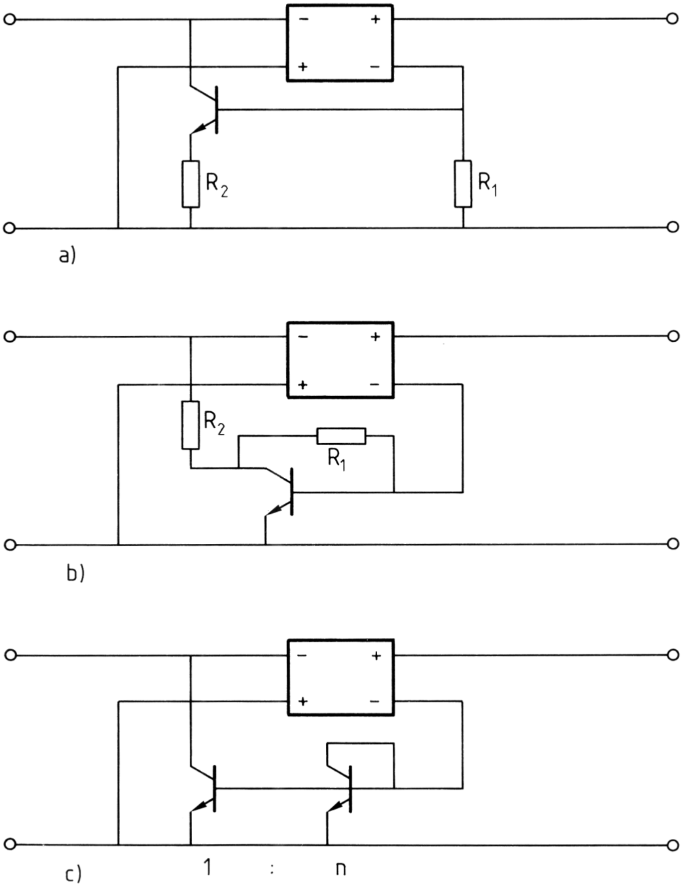

Figure 4.18. Three inverting current amplifiers with active feedback networks.

4.9 Discussion

Design aspects regarding linearity and accuracy were approached in this chapter in a rather unusual way. We introduced the differential-error-to-gain ratio (D-G ratio) as a figure of merit of amplifier stages in order to characterise their performance with respect to the above-mentioned quality aspects.

As a worst-case approach we assumed the differential errors of the individual transmission parameters of the CE stage to be uncorrelated and to have equal signs.

As expected, we found that the insertion of series and shunt impedances at the output of a stage leads to inferior performance, manifested by a relatively large increase in the D-G ratio, in relationship to their effect on output impedances. On these grounds brute-force methods for obtaining impedance adaptation at the output must be rejected.

The insertion of series and shunt impedances at the input of the output stage may change the character of the drive impedance so that the transfer of the stage is linearised, but it reduces at the same time the gain contribution of the stage to such an extent that the net result is negative; the D-G ratio increases.

The performance of local-feedback configurations was compared with that of the CE stage. Non-energic feedback, as present in the CC and CB stages hardly effects the D-G ratio. A relatively small increase in the D-G ratios results from local feedback with impedances in the feedback network (shunt stage and series stage).

Because local feedback leads to a reduction in at least one of the transfer parameters, it also results in an enhancement of the driving level needed to produce a certain output level. Because the driving stage has to handle larger signals, the application of local feedback to the output stage generally leads to greater non-linearity, unless compensations occur.

Compensations are considered to be *reliable,* only when they result from exactly opposite differential errors of corresponding parameters in two cascaded stages. This compensation technique is considered to be *useful* only when these cascaded stages behave as a single stage with respect to their highfrequency performance.

A practical conclusion is that for the highest accuracy and the lowest distortion (the number of stages being given), there should be no series or shunt impedances in the signal path, nor should local feedback he applied to the stages. If the exclusive use of CE stages cannot be tolerated, for example for reasons related to high-frequency performance, local feedback is preferred over the insertion of series and shunt impedances.

If the application of local feedback somewhere in the active part cannot be avoided, it should preferably be applied to the output stage because only in that case is the linearity of the loop gain improved, albeit at the expense of its magnitude. We will return to this subject in the next chapter. We must emphasise here that the design criteria found for optimum distortion and accuracy performance do not conflict with design criteria for optimum noise performance, at least not to a first-order approximation.

A conflict may arise in the case where the optimum bias current of the first stage with respect to noise performance is insufficiently large to supply the drive current for the second stage. Because the number of stages within one feedback loop cannot be increased without limit it may become necessary to use cascaded amplifiers in order to solve this design problem.

Quantitative considerations were not given in this chapter; they would carry us too far. We believe that they might be based fruitfully on the transmissionparameter approach as given here. Further work might involve the comparison of the performance of various types of active devices in the output stage. The available technology is a very dominant factor here, as it appeared to be for noise performance as well.

In this chapter we made allowance for the non-ideal isolation of the ports of the active part. The expressions for the asymptotic-gain values of configurations with series feedback at input and output were given. In both cases there is some inherent inaccuracy and non-linearity that can be combated by choosing types of stages such that the influence of the relevant transmission parameters is reduced.

We further examined in this chapter some inaccuracy and non-linearity aspects of indirect-feedback amplifiers. A condition for perfect linearity, which is however not practical can he found in these amplifiers, because it requires linear and accurately known source or load impedances.

The indirect-feedback technique might be judged to be exclusively suited to integrated circuit techniques because of the required accurate matching. This is true in the case of FET input and output stages. In the case of bipolar transistors, however, the required matching of the transconductances can be realised by using exactly equal bias currents. The application is therefore not reserved for integrated circuits exclusively.

A short qualitative discussion of active-feedback techniques led us to the conclusion that for good noise performance the active part of the feedback network must he located ‘close’ to the output, while it has to be close to the input for good distortion performance. A compromise will be necessary when both design aspects are important. Alternatively, the amplifier may he divided into two sections.

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5 Design Considerations Regarding Optimum High-Frequency Performance of Negative-Feedback Amplifiers

5.1 Introduction

In this chapter we will deal with some aspects of the high-frequency behaviour of *overall* negative-feedback amplifiers. These aspects, such as bandwidth, stability and accuracy of the high-frequency band edge, are closely related to the properties of the active devices and of source and load.

In chapters 3 and 4 we concluded that for optimum noise performance an input configuration, approximating the CE stage as closely as possible is preferred. For optimum performance with a fixed number of active devices in the feedback loop it was pointed out that CE stages are preferred elsewhere in the feedback loop as well. No strongly conflicting requirements for the structure of the active part of the feedback loop were found until now.

The question arises as to what extent the requirements on the amplifier-stage configurations to be used for optimum noise, accuracy and distortion performance are compatible with those for adequate high-frequency behaviour. The development of criteria for the capabilities of configurations in this respect is one of the subjects of this chapter.

The well-known gain-bandwidth product is not considered to be the most appropriate figure of merit for characterising the high-frequency-performance capabilities. Especially in cascaded stages its definition is rather artificial, and we will therefore introduce a somewhat different method to characterise the performance of amplifier stages.

The wish to realise specific singularity patterns of the transfer function A_t (such as MFM, MFD, etc.) implies the necessity for flexible design techniques for the pole-zero pattern of the loop transfer function (compensation techniques). Those compensation techniques that do interfere the least with design techniques for obtaining low noise, low distortion, etc. are preferable.

Design considerations will be given for *all-pole maximally-flat-magnitude* (MFM, Butterworth) filter characteristics exclusively, because the parameters of this MFM approximation of the ideal low-pass filter have simple relations to the amplifier bandwidth. The design procedure as outlined for this specific approximation is general, however, and can be followed for the realisation of other types of filter characteristics [1] as well.

5.2 Design Considerations Resulting from the Asymptotic-Gain Model

5.2.1 Introduction

For the design considerations regarding the frequency response of a negative-feedback amplifier we can advantageously use the asymptotic-gain model derived in chapter 1. The transfer function A_t of a negative-feedback amplifier can then be written as:

$$A_t = -A_{t\infty} \frac{A\beta}{1 - A\beta} \tag{5.1}$$

where the influence of the direct transfer ρ is ignored. Similar expressions are available for the input and output impedances of multiloop-feedback amplifiers. These were derived in chapter 1. The design considerations for accuracy etc. of these impedances are similar to those for the amplifier transfer function. The impedance functions and transfer functions are mutually related as was shown in chapter 1. This may necessitate compromises as to the accuracy of realisation of one or both functions. In this chapter we will deal exclusively with the amplifier transfer function.

The poles of the transfer function A_t are the poles of $A_{t\infty}$ and the roots of the equation $(1 - A\beta) = 0$. The zeros in A_t are given by the zeros in $A_{t\infty}$ and $A\beta$. As a matter of fact, the transfer function ρ can influence the location of the zeros. It is assumed, however, that this influence can be disregarded.

5.2.2 Bandwidth definition

Because $A_{t\infty}$ is the desired transfer function, the poles and zeros of $A_{t\infty}$ in the *midband frequency range* are intended to give the transmittance a prescribed frequency dependency. The bandwidth $B(\omega)$ of the transmittance can be defined as the difference between the high frequency and the low frequency for which the magnitude of the transfer function is reduced by 3 dB with respect to its intended value as specified by $A_{t\infty}$.

The bandwidth definition is thus disconnected from the intended frequency dependency of the amplifier transfer function. An example demonstrating the utility of this definition is given in figure 5.1, where the intended transfer function is that of an integrator with infinite bandwidth.

The high-frequency band edge (h.f. –3dB frequency) can be fixed either by the singularities of $A\beta/(1-A\beta)$ or by those of both $A_{t\infty}$ and $A\beta/(1-A\beta)$. In the following we will suppose that the bandwidth (as defined above) has to be maximised under the restraint that an MFM all-pole filtering has to be accomplished (in addition to the possibly required frequency dependency of $A_{t\infty}$, which will be left out of consideration from now on).

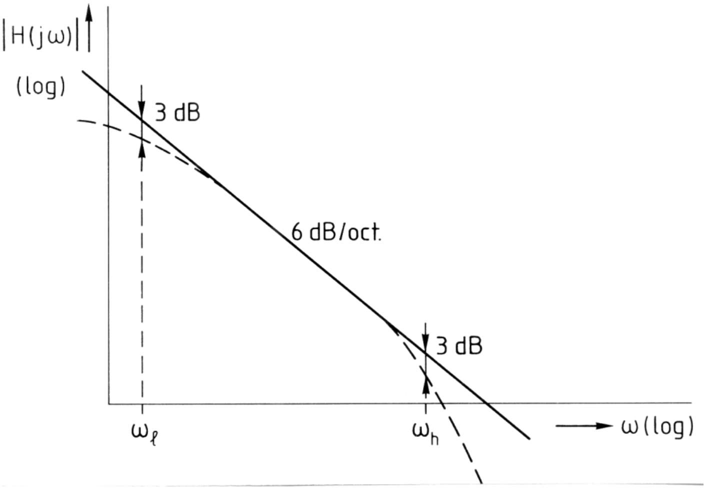


Figure 5.1 *Bandwidth definition for an integrator. Solid line: Intended transfer function $A_{t\infty}$. Dashed line: Actual transfer function A_t. $B(\omega) = \omega_h - \omega_\ell$.*

5.2.3 *Phantom zeros*

Because the factor β occurs in the expression for $A_{t\infty}$ (see chapter 1), a pole in $A_{t\infty}$ can be a zero of $A\beta$. Such zeros in $A\beta$ are called *phantom zeros* [2], because they do not occur in the transfer function A_t, yet have a large influence on the root locus, and hence, tend to fix the position of the high-frequency band edge.

This is illustrated in figure 5.2 by sketches of the root locus for a second-order feedback system first without and then with a phantom zero. Figure 5.2a shows the situation where no phantom zero is used. With increasing loop gain, the imaginary parts of the poles increase. A zero in the loop transfer function deflects the root-locus branches to the left so that damping is enlarged, as indicated in figure 5.2b. When this zero is a phantom zero, the pole-zero pattern of A_t exhibits only the two poles, because the zero in $A\beta$ is cancelled by the pole in $A_{t\infty}$ (figure 5.2c).

The zero tends to set the high-frequency band edge, all the more so when it is shifted further towards the poles. When it coincides with one of the poles, for example, the root locus degenerates into the negative real axis and the remaining pole is moved to the left. At the location where the pole and the phantom zero are fused, however, a pole in A_t remains, fixing the h.f. –3dB frequency.

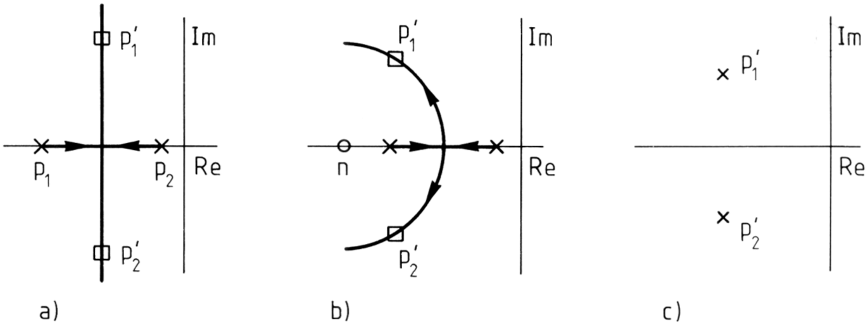


Figure 5.2 *Root locus of a second-order system. a. Without zero; b. with (phantom) zero; c. pole-zero pattern of A_t when n is a phantom zero.*

5.2.4 Relation between bandwidth and pole locations; the 'loop-gain-poles product'

The MFM all-pole filtering is chosen in this chapter as a vehicle for high-frequency-performance considerations, because the pole positions of the amplifier transfer function are simply related in that case to the bandwidth.

If the midband transfer function of the amplifier is denoted as $A_t(o)$ then the magnitude of the MFM transfer function can be written as:

$$|A_t(j\omega)| = \frac{A_t(o)}{\sqrt[k]{1 + \omega^{2k}/\omega_n^{2k}}} \tag{5.2}$$

The poles lie equally spaced on a semi-circle with radius ω_n which is centred in the origin [1]. The –3 dB bandwidth equals ω_n. The product of the poles is given by ω_n^k We can therefore write:

$$\omega_n = \sqrt[k]{\prod_1^k |p_i|}$$

where p_i are the poles of the transfer function A_t.

For the realisation of an all-pole MFM function, no zeros are tolerated in the loop transfer function $A\beta$ except phantom zeros. Thus AD can be written as:

$$A\beta = A_o\beta_o \frac{(1 - p/n_1)\ldots(1 - p/n_m)}{(1 - p/p_1)\ldots(1 - p/p_k)})$$

where $n_1 - n_m$ are phantom zeros, while $A\beta/(1 - A\beta)$ is given by:

$$\frac{A\beta}{1-A\beta}=\frac{A_o\beta_o}{1-A_o\beta_o}\;\frac{\left(1+\ldots p^m\prod_{1}^{m}\frac{1}{n_h}\right)}{\left(1+\ldots\frac{p^k}{(1-A_o\beta_o)}\prod_{1}^{k}\frac{1}{p_j}\right)}$$

The number of poles will always exceed the number of phantom zeros, and therefore, the following relation holds:

$$\prod_{i=1}^{k} p_i=(1-A_o\beta_o)\sum_{j=1}^{k} p_j$$

where p_i are poles of A_t, while p_j are the poles of $A\beta$.
If the poles p_i of A_t are located in Butterworth positions, then:

$$B(\omega)=\omega_n=\sqrt[k]{\prod_{i=1}^{k}|p_i|}=\sqrt[k]{|1-A_o\beta_o|\left|\prod_{j=1}^{k}p_j\right|}\qquad(5.3)$$

From this expression it can be seen that the attainable bandwidth of a negative-feedback amplifier is determined by the product of the low-frequency loop gain and the magnitudes of the poles of the loop transfer function, which will be referred to in the following as '*the loop-gain-poles product*' or the L-P product. In the case of an n-th-order MFM filter, the bandwidth equals the n-th root of the L-P product. In the case of other types of filter characteristics the bandwidth will have a less simple relation to the L-P product, but the latter will still be a measure for the realisable bandwidth.

The problem in designing so that maximum bandwidth is obtained is, of course, to force all the poles of $A\beta/(1-A\beta)$ into the desired positions (in this case Butterworth positions).

From the root-locus construction rules it is known that the directions of the asymptotes are determined by the difference between the number of poles and finite zeros. When this difference is larger than two, some of the asymptotes move away from the imaginary axis. The possibility of forcing all poles to lie on a circle centred at the origin may be hampered as a consequence.

5.2.5 Description of the design problem

If only a part of the poles can be manoeuvred into Butterworth positions, while the others are far removed from the origin in the left half-plane, the design can be considered as *less efficient*. In practice, though, this will always be the case, merely because an active circuit can be described only approximately as a network with finite numbers of network elements and singularities. Therefore, we have to resort to rather coarse approximations and to make a distinction between the *dominant* poles and the others.

Whether a pole is reckoned in the dominant group or not depends on the particular situation. Generally, we will refer to a pole of the loop transfer function as dominant if it can be located in the bandwidth-determining (Butterworth) group.

It must be noted here that in a negative-feedback amplifier the excess phase of the non-dominant poles can endanger the stability of an amplifier. More detailed information about the influence of non-dominant poles is given at the end of this chapter.

The design problem can be formulated now as follows:

1. The product of low-frequency loop gain $A_o\beta_o$ and the moduli of the dominant poles of the loop transfer function must be made as large as possible.
2. These dominant poles must have positions such <that the poles of the transfer function A_t can be forced into the desired positions with the aid of suitable *compensation* techniques.

It doesn't make sense, of course, to try and increase the L-P product without limit because, as formulated previously, some branches of the root locus move away from the imaginary axis while others move into the right half-plane.

We will, for the time being, approach this design problem without paying attention to the other transfer-quality aspects. For a given type and magnitude of the desired transfer function and for given source and load impedances the optimum amplifier-stage configurations with respect to *bandwidth capabilities* must then be found. We will therefore investigate in the next section the bandwidth capabilities of the single-device configurations, which are the available building blocks of the active part. Later on we will again consider noise, distortion and accuracy aspects.

5.3 Characterisation of the Bandwidth Capabilities of Single-Active-Device Configurations

A measure for the bandwidth capabilities of single-active-device configurations inserted in the active part of an overall negative-feedback amplifier is their contribution to the product of the loop gain and magnitudes of the poles (L-P product). Because the loop gain is dimensionless and because it refers to a closed loop, it is immaterial — for finding this contribution — whether the current or voltage transfer function of a stage is considered.

Bandwidth capabilities of amplifier stages are somehow related to the current-gain-

factor transit frequency ω_T of the active devices. The transit frequency can be regarded as a figure of merit of the device. It can be expressed in the parameters of the equivalent circuits for both transistors and FET's as introduced in chapter 2.
In the case of a bipolar transistor:

$$\omega_T = \frac{g_e}{C_{b'e} + C_{b'c}}$$

and for a field-effect transistor:

$$\omega_T = \frac{g_m}{C_{gd} + C_{gs}} = \frac{g_m}{C_{iss}}$$

Frequently, these figures of merit are indicated as gain-bandwidth products, being the products of the low-frequency current-gain factor and its –3 dB bandwidth. They thus relate to the rather artificial particular amplifier circuits given in figure 5.3.

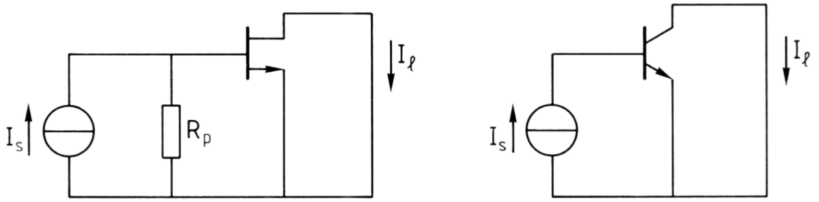


Figure 5.3 *Circuit configurations in which the product of the low-frequency current gain and –3 dB bandwidth equals ω_T.*

The transfer functions I_ℓ/I_s in figure 5.3 comprise one pole. The product of low-frequency current gain and the magnitude of this pole is therefore equal to the product of the gain and bandwidth. That the term 'gain-bandwidth product', however, is not appropriate and even confusing for cases where more poles are involved, may be illustrated by the following example.

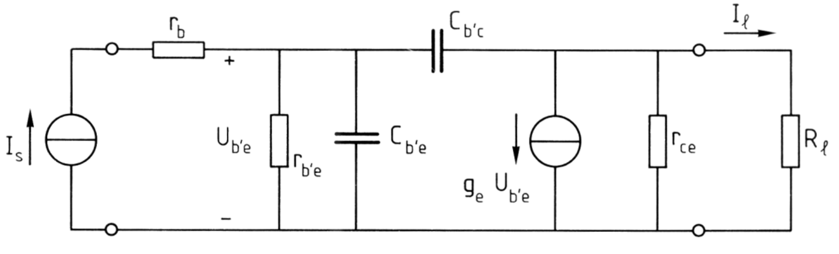


Figure 5.4 *Equivalent circuit of a CE configuration loaded with a resistance.*

The product of the current gain and bandwidth in a CE configuration can be considerably smaller than ω_T when the stage is loaded with a non-zero resistance. This situation is shown in figure 5.4 where the transistor is represented by its hybrid-π equivalent circuit. The low-frequency current gain of this stage is smaller than its current-gain factor because of the current division at the output between R_ℓ and r_{ce}. The bandwidth is reduced as well as a consequence of the enlarged influence of $C_{b'c}$, commonly known as the Miller effect.

It is possible to introduce a new ‘CB product’ [3] related to ω_T in the following way:

$$CB' = \frac{\omega_T}{1 + \omega_T R_\ell C_{b'c}} \tag{5.4}$$

This ‘CB product’, however, is the product of the low-frequency current gain (neglecting the influence of r_{ce}) and the magnitude of the *dominant* pole. The current gain has a second pole. This pole may have a hardly noticeable influence on the bandwidth of the stage, but it cannot straight away be disregarded for the characterisation of bandwidth capabilities of such a stage in a negative-feedback amplifier, because it may arrive in the dominant group as a consequence of the negative-feedback action. Therefore, expression (5.4) is misleading as a figure of merit for a stage in the active part of an overall negative-feedback amplifier.

In the above case of a real load impedance, the second pole may admittedly be very large and will generally not arrive in the dominant group. The contribution to the L-P product is adequately described then by the ‘gain-bandwidth product’ as given in (5.4). However, if the load is formed by an R_ℓ-C_ℓ combination, which is a more usual situation, the dominant pole has the same value as in the resistance-loaded case as long as $C_\ell < \alpha_{eo} C_{b'c}$. The second pole can easily enter into the dominant group. Besides, there is a zero in the left half-plane.

The transfer function is given by:

$$A_i = \frac{I_\ell}{I_s} = -\frac{(g_e - pC_{b'c})(g_\ell + PC_\ell)}{(g_{b'e} + PC_{b'e})(g_\ell + pC_\ell + PC_{b'c}) + pC_{b'c}(g_\ell + PC_\ell + g_e)}$$

The right-half-plane zero introduces some excess phase; generally it can be ignored.

The product of the poles is found to have a value:

$$p_1 p_2 \approx \frac{w_T}{\alpha_{eo}} \frac{1}{R_\ell C_\ell}$$

The low-frequency value A_{io} of the current gain is approximately equal to the low-frequency current-gain factor α_{eo} of the transistor. Because the left-half-plane zero $n = -1/R_\ell C_\ell$ can be cancelled by a pole elsewhere in the loop, the *effective* contribution to the L-P product of this stage can be given as:

$$A_{io}\left|\frac{p_1p_2}{n}\right|\approx\omega_T \tag{5.5}$$

No reduction in the contribution to the L-P product — due to the Miller-effect — is observed. The ‘gain-bandwidth product’ as given in (5.4) is obviously not an adequate figure of merit for bandwidth capabilities in this situation.

We can conclude that the contribution to the L-P product is a more adequate measure for characterising the capabilities of configurations in negative-feedback amplifiers. This contribution can equal the transit-frequency ω_t of the device in the case of an RC-loaded, current-driven CE stage, as we have demonstrated above.

5.4 Comparison of the Single-Device Configurations

5.4.1 Introduction

In previous chapters we compared the merits of the various single-device configurations in combination with shunt and series impedances with respect to performance aspects pertaining to noise, distortion and accuracy. We found that series and shunt impedances are to be avoided and that non-feedback stages are preferably used in the active part of a negative-feedback amplifier.

In order to judge in this section the single-device configurations with respect to their high-frequency-performance capabilities when used in the active part of an overall negative-feedback amplifier, we now will similarly compare their contributions to the L-P product for various source and load conditions. The considerations are given mainly for bipolar transistors, but the results are general and can be equally well applied to junction FET’s unless explicitly stated otherwise.

5.4.2 High-frequency performance of CE, CC and CB stages

In the previous section, we saw that a current-driven and RC-loaded CE stage gives an effective contribution of ω_T to the L-P product. Equally large contributions are obtained with a CC stage and a CB stage under the same drive and load conditions.

The poles of the CC stage are even in the same positions under the above drive and load conditions as in the CE stage, because the transfer is determined by the transimpedance ζ and the current-gain factor α, which are not affected by feedback. Alternatively, we can say that there is no local loop gain in a current-driven CC stage via paths other than $C_{b'c}$. It may be clarifying to look at the equivalent circuit of the CC-stage-feedback loop, which is depicted in figure 5.5.

Low-frequency loop gain is reduced to zero, the same as in a CE stage, when Z_s is made infinitely large. The transfer function I_ℓ/I_s has a zero at $n = -\omega_T$, which will generally have a beneficial influence on stability.

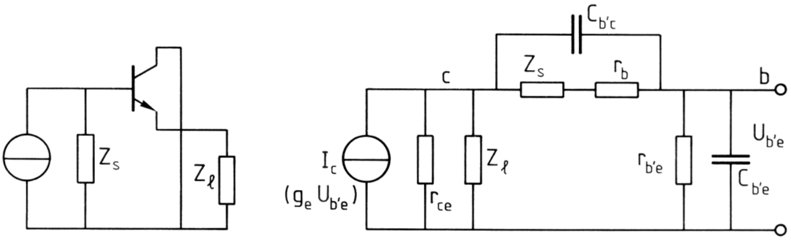


Figure 5.5 CC stage with equivalent circuit of its feedback loop.

The contribution of an RC-loaded CB stage to the L-P product also equals ω_T, but now the poles are in the neighbourhood of $-\omega_T$ and $-1/R_\ell C_\ell$, respectively. The second pole coincides with the zero in $-1/R_\ell C_\ell$. The resulting pole at $-\omega_T$ will frequently not belong to the dominant group.

It is important to note that for this reason a CB stage can frequently be used as a *buffer stage* in the active part without having much influence on the L-P product. In section 5.5 we will return to this subject.

5.4.3 The influence of series and shunt resistances on the L-P product

In chapters 3 and 4 we found that the insertion of shunt and series resistances in the active part of an amplifier can seriously harm the noise, distortion and accuracy performance. Nevertheless, it seems to be useful to investigate their influence on the contribution of a stage to the L-P product, if only because the use of relatively small shunt resistances is common practice in wide-band amplifiers.

First we will examine their influence in the case of the CE stage and the non-energic local-feedback stages. Then, in the next section, we will deal with the high-frequency performance of the impedance feedback stages again including series and shunt resistances.

(i) CE stage

Figure 5.6a and b depicts two relevant situations for CE stages with a series output-resistance and a shunt input resistance, respectively.

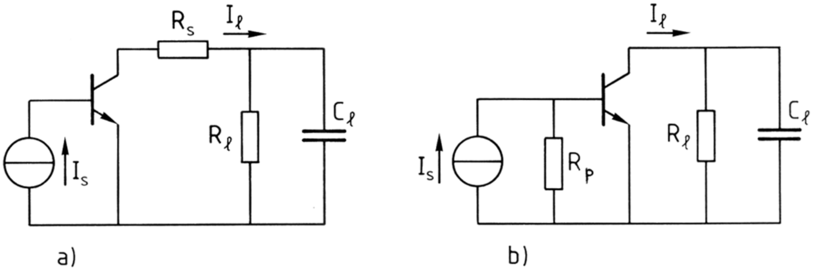


Figure 5.6 *CE stages with series resistance at the output and shunt resistance at the input, respectively.*

In the case of figure 5.6a, there are three poles, two of which very likely belong to the dominant group. If we take into account only these two poles, we find a reduction in the effective contribution to the L-P product:

$$\left|\frac{A_i \prod p}{n}\right| = \frac{\omega_T}{1 + \omega_T C_{b'c} R_s + \frac{C_{b'c}}{C_\ell}\left(1 + \frac{R_s}{R_\ell}\right)}$$

A similar approximation for figure 5.6b leads to a contribution:

$$\left|\frac{A_i \prod p}{n}\right| = \frac{\omega_T}{1 + r_b/R_p}$$

The situation of figure 5.6a corresponds to an approximation of a practical situation where, for example, R_s and R_ℓ-R$_\ell$ form the input impedance of a second CE stage which in turn has a low load impedance. In that case, the resistance R_s corresponds to r_b. In a cascade of CE stages a small reduction as to ω_T in the contribution to the L-P product therefore occurs.

The situation of figure 5.6b is representative for so-called resistive broadbanding (to be dealt with in section 5.6.2) and for input stages in amplifiers for resistive signal sources. Obviously, small shunt resistances may lead to a strong reduction in bandwidth capabilities in the case of bipolar-transistor stages. This reduction can be nullified by inserting an inductance in series with R_p, as shown in figure 5.7. Of course, this is not feasible in the case of a resistive signal source. The inductance can be chosen so that a pole-zero cancellation is accomplished. It can be used also for shunt peaking, which means that complex poles and an additional zero are generated. In both cases, the effective contribution to the L-P product amounts to ω_T.

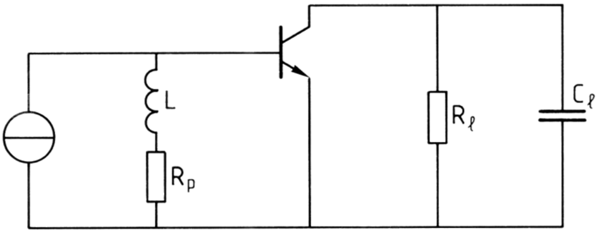


Figure 5.7 Compensation for loss of bandwidth capabilities.

Shunt resistances at the input of FET stages do not reduce the contribution to the L-P product, because of the absence of a series input resistance.

(ii) CC stage

A shunt resistance at the input of a CC stage leads to an equal reduction in the L-P-product contribution as in the CE stage. It can he nullified here as well with a series inductance. In the case of the CC stage, however, the shunt resistance has a much more violent effect on the pole locations because the feedback loop is closed and a strong interaction between the poles is involved. From figure 5.5 it can he seen that the loop transfer function has two poles and a zero. This zero $n = -1/(r_p + r_b)C_{b'c}$ is a *phantom zero,* because $A_{u\infty}$ appears to have a pole at the same place. The p-n pattern of the loop transfer function and the magnitude of the loop gain easily give rise to complex poles in the stage transfer function. The insertion of a series inductance tends to yield larger imaginary parts of these poles. Even a Colpitt oscillator may result! The technique is not very attractive for use in negative-feedback amplifiers. With a real load resistance, however, one pole may lie outside the dominant group and it would be possible to use the series inductance to compensate for the reductions caused by $C_{b'c}$, comparable to the Miller-effect reduction in the CE stage (figure 5.3). The non-dominant pole can he forced into the dominant group by this technique.

A series resistance at the output of a CC stage has a similar effect on the L-P product as in the case of a CE stage.

(iii) CB stage

A shunt resistance at the input of a CB stage leads to a somewhat larger reduction in its L-P product contribution. It is given by:

$$\left|\frac{A_i \prod p}{n}\right| = \frac{\omega_T}{1 + r_b/R_p + \omega_T r_b C_{b'c}(1 + r_b/R_p) + r_b C_{b'c}/R_\ell C_\ell}$$

The last two terms of the denominator can, however, generally be disregarded. The reduction can he nullified here as well, as far as the factor r_b/R_p in the denominator is concerned, by inserting a series inductance.

A series resistance at the output of a CB stage has similar effects on the L-P product as in the case of a CE stage. This reduction, however, is of minor significance because it refers to a pole that will generally lie outside the dominant group.

5.4.4 High-frequency performance of shunt and series stages

The other single-device configurations all use impedances for realising the desired feedback action. When resistances are used, these act similar to series and shunt resistances. As a result, their contributions to the L-P product are inherently smaller than those of the CE stage and the non-energic feedback configurations. Only the shunt stage and the series stage will be examined in some detail.

(i) Shunt stage

The feedback resistance of a shunt stage will generally be fairly large compared to the base resistance r_b. The reduction of its L-P product contribution will therefore be small. For real load impedances, an inductance can be taken in series with the feedback resistance, in order to compensate for the reduction due to $C_{b'c}$ (Miller effect).

Note that the equivalent circuit of the loop of the shunt stage and the CC stage are equal. The feedback impedance of the shunt stage plays the same role as the source impedance of the CC stage.

(ii) Series stage

The reduction in the contribution to the L-P product in series stages can be substantial even when compensation techniques are used. For a well-behaved transfer function of a series stage, a shunt capacitance across the feedback resistance is desirable with a value $C_E = 1/\omega_T R_E$ (emitter compensation) [4].

Figure 5.8 shows a current-driven and RC-loaded series stage where this compensation is applied. The transmission parameters of this stage have the same form as those of the CE stage (see Table 2.4). In spite of the compensation capacitor, the contribution to the L-P product is decreased and amounts to:

$$\left|\frac{A_i \prod p}{n}\right| = \frac{\omega_T}{1 + \omega_T R_E C_{b'c}}$$

The emitter series resistance appears to have, as expected, an adverse influence similar to that of a series resistance in the collector lead. It is possible of course to insert a larger shunt capacitance across the emitter resistance. The pole-zero cancellation making the stage well-behaved is no longer present in that case.

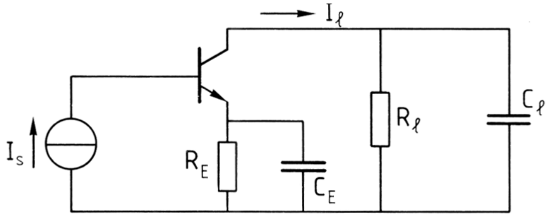


Figure 5.8 Series stage with compensation capacitance C_E for obtaining a well-behaved transfer function.

A further reduction in the L-P product contribution is brought about when a shunt resistance R_p is inserted at the input. The reduction factors equals $(1 + r_b/R_p)$, which is the same value as in a CE stage. This reduction factor can *also be nullified* here with the aid of an inductance in series with R_p.

5.4.5 Influence of shunt capacitances

We have dealt with shunt and series resistances in the signal path and methods for overcoming the associated reduction in the contributions to the L-P product. The influence of shunt capacitances was not yet explicitly examined. It goes without saying, however, that shunt capacitances in the signal path generally deteriorate high-frequency performance. Capacitances that are small compared to the inherently present capacitances may be added without reducing bandwidth capabilities seriously. In order to accomplish substantial effects on the p-n pattern of the loop transfer function these capacitances must be part of a feedback network. They can he inserted, for example, between the source and load in a CE stage in order to obtain ‘pole splitting’. In section 5.6 we will return to this subject.

5.4.6 Conclusion

At the end of this section we can draw some interesting conclusions.

First, there appears to be no need for using small shunt resistances in the signal path of a wide-band amplifier. Though the insertion of such resistances may seem obvious in order to give the poles large magnitudes, they have an adverse effect on the bandwidth capabilities of a stage because the loop gain is reduced more than proportionally. The contribution to the L-P product of a bipolar-transistor stage with an input shunt resistance is therefore less than optimal.

Series resistances, when inserted at the input of a stage, were not examined separately as their influence is manifest in the same way as that of r_b.

Series resistances at the output of a stage adversely influence its contribution to the L-P product as well.

Apart from the apparent detrimental effect of series and shunt resistances, we can further conclude that the L-P product hardly depends at all on the types of single-device configurations used in the active part of a negative-feedback amplifier. Only the series stage is somewhat inferior to the other single-loop stages.

The equivalence of the stages does not imply that, from the point of view of high-frequency performance, the types of configurations can be chosen arbitrarily because a second important design aspect has to he taken into account. The poles of the amplifier transfer function need to be manoeuvred into the desired (Butterworth) positions.

Up to now we have not found any design criteria for optimum high-frequency performance that conflict with the criteria we found earlier for optimum performance with regard to other quality aspects. Consequently, preference for certain stage configurations do arise, for the time being, from considerations regarding noise, accuracy and linearity.

5.5 Designability of the Desired Pole-Zero Pattern of a Loop Transfer Function

5.5.1 Introduction

An additional important design aspect that will he considered now is the *designability* of an appropriate singularity pattern of the loop transfer function $A\beta$. This is one aspect that may give rise to a preference for stages other than the CE stage in the feedback loop.

If the use of such other stages becomes imperative, it will be necessary to judge these with respect to their performance on the basis of other quality aspects. Naturally, those stage configurations that harm the noise and distortion performance the least, will be preferable.

5.5.2 Designability of cascaded CE stages

A cascade of CE stages is, on the grounds of the criteria found thus far, the most obvious structure of the active part of a negative-feedback amplifier. In a cascade of CE stages, however, there is a rather strong interaction between the transfer functions of the individual stages, because all transfer-parameters play a significant role. We will illustrate this by an example.

Figure 5.9 shows a cascade of two CE stages with an RC load. To calculate the first-stage transfer function the input impedance of the second stage must be known. A calculation of this impedance using the hybrid-n equivalent circuit yields:

$$Z_{iT_2} = r_b + \frac{Y_{b'c} + Y_\ell}{Y_{b'e}(Y_{b'c} + Y_\ell) + Y_{b'c}(Y_\ell + g_e)} \tag{5.6}$$

where $Y_{b'c} = pC_{b'c}$, $Y_\ell = 1/R_\ell + pC_\ell$, $Y_{b'e} = 1/r_{b'e} + pC_{b'e}$.

This impedance can be decomposed into an equivalent network as given in figure 5.10 [5].

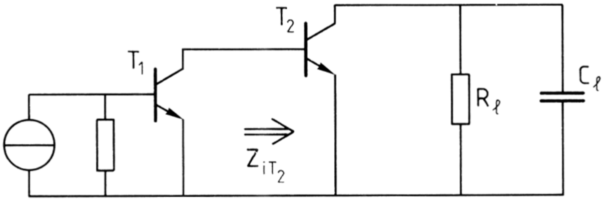


Figure 5.9 Two cascaded CE stages with an RC load.

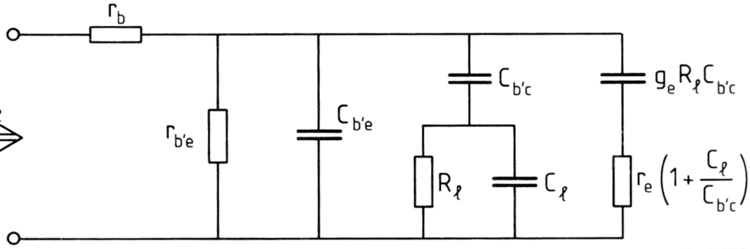


Figure 5.10 Equivalent circuit for the input impedance of a CE stage with an RC load.

The input impedance of this rather complicated network comprises two poles and two zeros. It constitutes the load impedance of the first transistor. This renders the transfer function of the first stage difficult to design. Because of the interaction between the load and input impedance the cascaded CE stages are not very *designable*. In order to avoid this strong interaction and to improve the designability a situation where the transfer of a stage is determined to a large extent by one of its transfer parameters is preferably realised. This can be achieved in various ways.

One method for reducing the interaction is to add series or shunt resistances to a CE stage in order to have the transfer function determined by essentially one transfer parameter. It will be clear from previous considerations that various quality aspects are harmed by such design measures. Nevertheless, it has been a standard design technique from the very beginning of wide-band-amplifier design to use low shunt resistances (often referred to as load resistances) in the signal path for so-called resistive broadbanding.

5.5.3 Designability of cascaded local-feedback stages

A very powerful technique for avoiding strong interaction with much less degradation in quality can be obtained by the use of local-feedback stages in the feedback loop. The loop transfer can then be determined to a large extent by the product of individual transfer parameters, the singularities of which are known from chapter 2.

By combining the proper types of configurations, very designable transfer functions can be obtained. Some combinations are discussed below:

- The application of shunt stages and CC stages in combination with CE stages easily leads to strong interaction and to complex poles in the loop gain. These combinations may be useful for various purposes but they are seldom suitable for use in overall negative-feedback amplifiers.
- The combination of series stages with other configurations leads to suboptimum contributions to the product of the poles and loop gain, because of the increased influence of $C_{b'c}$.
- The combination of a CE stage and a CB stage in the form of a so-called *cascode* (depicted in figure 5.11) is well behaved, just as the combination of a CC stage and a CB stage, depicted in figure 5.12, which is frequently referred to as a *long-tailed pair*.

The cascode *(inverting)* and the long-tailed pair *(non-inverting)* will be discussed in more detail because of their great significance.

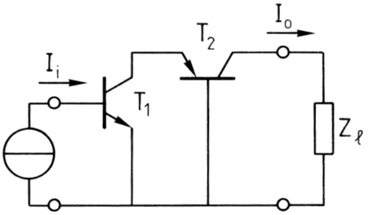


Figure 5.11 Cascade of CE and CB stage; also known as cascode (inverting).

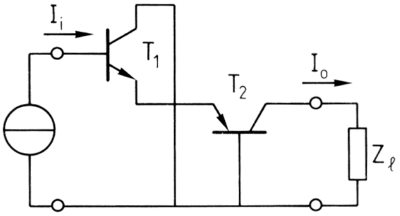


Figure 5.12 Cascade of CC and CB stage; also known as long-tailed pair (non-inverting).

(i) The cascode configuration

In a cascode configuration, the CE stage is loaded with the input impedance of a CB stage which is low provided that the loop gain in this current follower is large. Referring to figures 2.5 and 2.6 it can be seen that the loop gain can be large as long as the load impedance of the CB stage remains small with respect to r_{ce}. Because of the low load impedance, the CE stage exhibits a low voltage gain and the effect of $C_{b'c}$ on the dominant-pole position is virtually eliminated.

In a cascade of such stages there is virtually no interaction and, as each stage is driven from a very-high-impedance source (except perhaps the first stage), the influence of r_b on bandwidth capabilities is eliminated. The transfer function of the cascode is determined almost entirely by the product of the transfer parameters a of the individual stages. It can be approximated by:

$$\frac{I_o}{I_i} \approx -\frac{\alpha_{eo1}}{1 + \mathrm{j}\omega\alpha_{eo}/\omega_{T1}} \cdot \frac{\alpha_{bo2}}{1 + \mathrm{j}\omega/\omega_{T2}}$$

where the zeros in the transfer functions are ignored and where α_{bo} is the low-frequency current-gain factor of the CB stage. It has a value very close to unity.

A cascode stage thus has a dominant pole in $p = -\omega_{T1}/\alpha_{eo}$ and a far-removed pole in $-\omega_{T2}$, when it is current driven and loaded with an impedance which is small with respect to r_{ce}. The pole in $-\omega_{T2}$ will seldom belong to the dominant group. Only in the case of very large bandwidth must its influence be taken into account. The combination of CE and CB stages can therefore generally be regarded as *one stage*.

Since the singularities of the CB stage transfer function in the cascode configuration are frequently not used to enlarge the product of the dominant poles and loop gain, the addition of the CB stage may lead in a way to a less efficient design as far as the number of devices is concerned. We do not consider this to be a drawback, because the number of active devices is becoming an ever less important cost factor in many applications.

(ii) The long-tailed-pair configuration

The long-tailed pair of figure 5.12 is nearly equivalent to the cascode and an equally well-behaved combination. To appreciate this it may be clarifying to look first a little bit closer at the CC stage and its feedback loop as given in figure 5.5. If the CC stage is driven from a high-impedance source such as a cascode configuration, the low-frequency loop gain is virtually reduced to zero. For non-zero values of Z_ℓ, there is only the parasitic feedback due to $C_{b'c}$ shunted by the output capacitance of the driving cascode configuration. The dominant pole is reduced in a way similar to that in a CE stage (Miller effect) but the reduction is determined in this case by the shunt capacitance at the input as well.

The capacitive feedback can obviously be eliminated by virtually short circuiting the output. In the case of the long-tailed pair this is accomplished by the low input impedance of the CB stage which represents Z_ℓ in Figure 5.5.

When current driven and loaded with a relatively low impedance, the transfer function of the long-tailed pair is largely determined by the product of the transfer parameters α of the transistors, the same as with a cascode, as follows from the considerations above.

The pole at $-\omega_T$, contributed by the CB stage, is cancelled now by a zero in the transfer of the CC stage (see chapter 2) and, apart from the transmission zeros of the CB stage which can generally be neglected, the transfer I_o/I_i is given by:

$$\frac{I_o}{I_i} \approx \frac{\alpha_{eo}}{1 + \mathrm{j}\omega\alpha_{eo}/\omega_T}$$

The input impedance of the long-tailed pair is — when bias currents in both stages are equal — about twice as large as that of the cascode configuration, and the transfer is somewhat more sensitive to parasitic shunt impedance at the input. Its noise performance is inferior to that of the cascode, because T_1 is a local-feedback configuration (see chapter 3). It should not be used as an input stage for this reason.

5.5.4 Conclusion

The considerations given above regarding the designability of the loop transfer function of a negative-feedback amplifier may have made it clear that it is an enormous advantage to have at one's disposal nearly *unilateral* stages. This desirable property was found in the inverting cascode stage as well as in the non-inverting long-tailed pair, which behave — when cascaded — as unilateral current-amplifier stages.

Of course, it is not *necessary* to use such configurations — which behave as single stages as far as their contribution to the L-P product are concerned — everywhere in the active part. Contributions to the L-P product equal or nearly equal to ω_T can be obtained with most of the single-device configurations as well. The main function of the CB stages is that they act as *buffers*, while the number of dominant poles is not necessarily influenced by their insertion.

In a cascade of the current-amplifier stages described, poles of comparable magnitudes occur in $-\omega_T/\alpha_{eo}$. If the active part were to be set up as such a cascade, this would frequently yield a pole-zero pattern of the loop transfer function $A\beta$ not suitable for realising an MFM characteristic of the amplifier transfer function A_t. Therefore, it will be necessary to take measures in order to force the root-locus branches to arrive at the desired pole positions.

These measures may involve the abandonment of the cascode and the long-tailed pair. The use of other single-device configurations can then be considered for obtaining a *controlled interaction* of the poles. The techniques available for this so-called *compensation* will be dealt with in the next section.

5.6 Methods to Force the Root Locus Through the Desired Pole Positions of A_t (Compensation Techniques)

5.6.1 Introduction

Bandwidth capabilities of the active devices, i.e. their contributions to the L-P product, can only fully be utilised if the number of stages and thus the number of poles is restricted. If the pole-zero pattern of $A\beta$ comprises three or more poles more than zeros, at least one branch of the root locus moves away into the left half-plane and at least one pole may become non-dominant in the transfer function A_t. If it does, bandwidth capabilities cannot be fully employed.

No zeros other than phantom zeros are allowed to influence the loop transfer function AD in the relevant frequency band, because otherwise they would undesirably occur in A_t as well.

The use of phantom zeros means that the attenuation in the passive part of the feedback loop is reduced beyond the frequencies corresponding to these zeros. Possibilities to realise them are not always available. When they are not, one has to resort to such a manipulation of the poles of the loop transfer function that the poles of A_t arrive in the correct positions. These possibilities are restricted as well, as will be shown in future sections.

The poles can be shifted to the required positions by using compensation techniques. We will discuss some of these techniques in the following. The discussion will concern unilateral stages, the transfer function of which is changed so that one or two pole locations are altered without changing the contribution to the L-P product, if possible. The techniques may conflict with design requirements regarding noise and linearity. The qualitative influence on these quality aspects will be discussed.

5.6.2 Resistive broadbanding

A pole in the loop transfer function, determined by a current-driven impedance consisting of a resistance in parallel with a capacitance, can easily be enlarged by inserting an additional shunt resistance. Apart from the possibility to shift a pole in this way, which may be desirable for manipulating the pole-zero pattern, various quality aspects may be, as we have argued earlier in various places, seriously harmed.

Though this method for changing the pole-zero pattern has large disadvantages and cannot be recommended, we will give an example of its application at the input of a current-driven cascode stage. Figure 5.13 shows the situations to be compared together with the associated Bode plots.

The pole in the current transfer function I_o/I_s is shifted by the shunt resistance. The contribution to the loop gain is reduced more than proportionally in the case of a bipolar transistor because of r_b, as we have seen in section 5.4.3, thereby reducing the contribution to the L-P product. As a reminder we note that the input noise current of the stage is enlarged by the shunt resistance and that the D-G ratio is enlarged as well.

By taking an inductance in series with R, the loss in the L-P product contribution can be avoided as was mentioned before in section 5.4.

In the case of an FET stage, there is no reduction in the contribution to the L-P product. The other disadvantages, however, are similar to those for the bipolar transistor.

Somewhat aside, it is perhaps worth noting that resistive broadbanding of a stage in an overall negative-feedback amplifier is from the viewpoint of bandwidth capabilities not as inefficient as it is in multistage cascades (bandwidth narrowing).

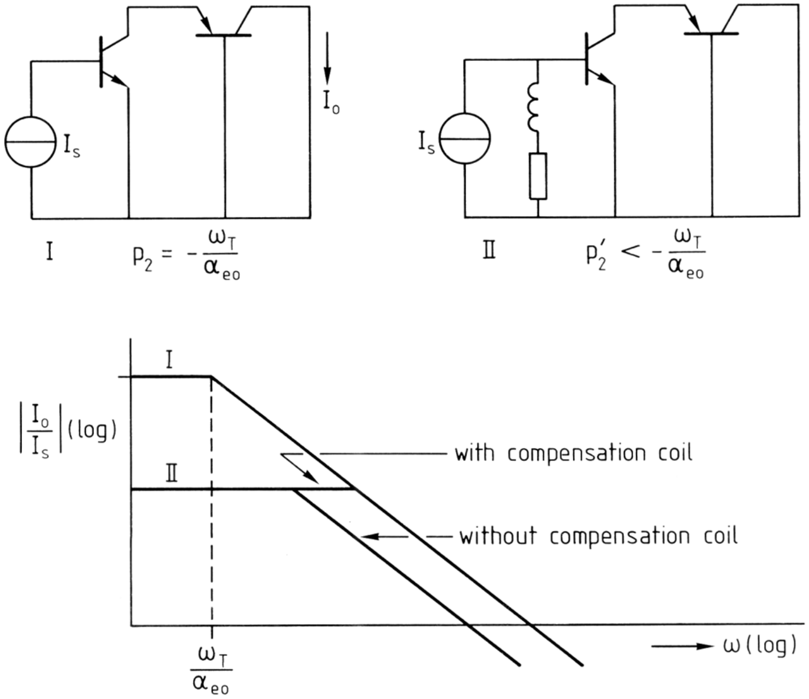

I_o
I_s
I
$p_2 = -\frac{\omega_T}{\alpha_{eo}}$
I_o
I_s
II
$p_2' < -\frac{\omega_T}{\alpha_{eo}}$
$\left|\frac{I_o}{I_s}\right|$ (log)
I
II
with compensation coil
without compensation coil
ω (log)
$\frac{\omega_T}{\alpha_{eo}}$

Figure 5.13 Example of resistive broadbanding in a cascode stage.

5.6.3 Pole splitting resulting from pole-zero cancellation

A compensation technique somewhat less harmful than resistive broadbanding uses a capacitor in series with the broadbanding resistor, as shown in figure 5.14.

The associated Bode plots are given as well. The influence of the shunt impedance on the magnitude and linearity of the loop gain, for example, can at least be small at low frequencies. A pole and a zero result from the insertion of the series capacitance. The zero must be cancelled by a pole elsewhere in the loop. Consequently, the original poles disappear and the resulting poles are split.

A second example of pole-zero cancellation for the realisation of pole splitting is given in figure 5.15. A Bode plot similar to that of figure 5.14 is obtained in this case where a cascade of a non-inverting stage with two feedback loops (see figure 2.4g) and a CB stage is used. In the first stage the current-gain factor α is fixed and reduced by the local feedback.

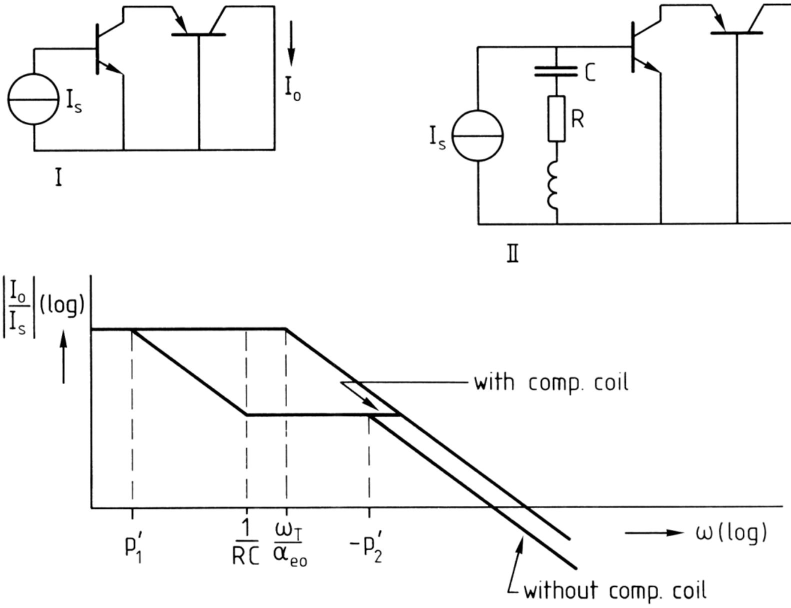


Figure 5.14 Pole-zero cancellation in a cascode stage.

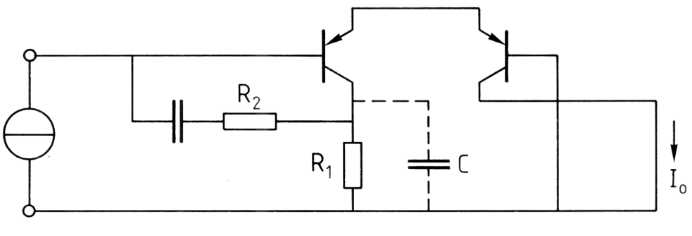


Figure 5.15 Non-inverting amplifier stage providing a transfer function with two poles and a zero.

The quality will be less harmed now, because the pole splitting is obtained by local feedback, allowing a larger resistance and smaller capacitance for a result similar to that of figure 5.14. Of course, $C_{b'c}$ has a certain influence on the transfer function. The reduction in the contribution to the L-P product due to the series resistance R_1 can be counteracted by a capacitor across R_1 (dashes in figure 5.15).

5.6.4 Pole splitting by means of local capacitive feedback

(i) Pole splitting in a GE stage

The interaction between load impedance and input impedance caused by $C_{b'c}$ in a CE stage can intentionally be used for pole splitting if the load impedance behaves as an RC load.

The enlargement of the input capacitance caused by the effect of the feedback capacitance with a factor $(1 - A_u)$ is generally known as the Miller effect. For a real load impedance such a description is usually sufficiently accurate. The second pole in the transfer function of the stage is very large in that case. In the case of an RC load, the Miller-effect description can also be used for finding the dominant pole as long as $C_{b'c} > C_\ell/\alpha_{eo}$.

In order to find the second pole, the mere introduction of an enlarged input capacitance is not adequate. The poles of the transfer impedance must then be calculated from (5.6) (section 5.5.2), or can be estimated from the equivalent circuit of figure 5.10, because these poles are the same as those of the input impedance.

A unilateral cascode and a CE stage with RC loads are compared in figure 5.16. The transimpedance of the cascode stage is (neglecting very large poles and zeros) given by:

$$\frac{U_\ell}{I_s} = \frac{-\alpha_{eo}R_\ell}{\{1 + p\alpha_{eo}/\omega_T\}\{1 + PR_\ell(C_\ell + C_{b'c})\}}$$

The transimpedance of the CE stage can, ignoring the coefficient of p, be written as:

$$\frac{U_\ell}{I_s} = \frac{-\alpha_{eo}R_\ell}{1 + p(\ldots) + p^2R_\ell(C_\ell + C_{b'c})\alpha_{eo}/\omega_T}$$

The product of the poles is nearly the same in both cases. In the CE stage the sum of the poles is enlarged by the capacitive feedback via $C_{b'c}$ and the poles are split apart. With the aid of an external capacitor between collector and base, the sum can be enlarged further. The product of the poles is somewhat reduced in that case.

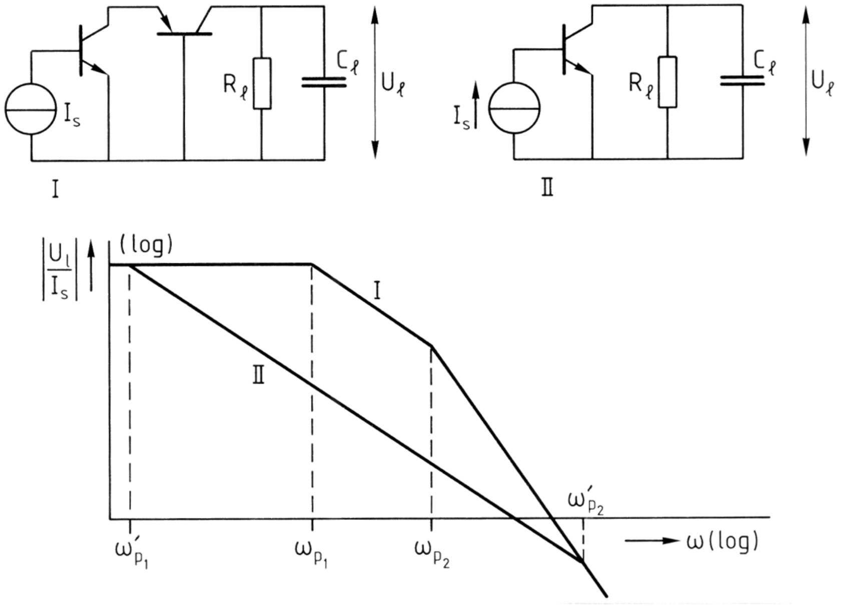


Figure 5.16 Current-driven cascode with EG load (I) compared with current-driven CE stage with EG load (II) with Bode plots of the transimpedances.

It is sometimes possible as an alternative to enlarge the load impedance and thus increase the effect of $C_{b'c}$. This possibility exists when the load is formed by the input impedance of a second stage. By selecting a series stage instead of a CE stage for the second stage, the desired effect may be obtained. Here also, a small reduction in the L-P product may result.

(ii) Pole splitting in a CC stage

The same pole-splitting effect occurs in the current-driven CC stage as shown in figure 5.17 and compared now with the non-inverting current-amplifying longtailed pair.

In this case, shunt capacitance across the input terminals adds to the feedback capacitance $C_{b'c}$, and the pole splitting is more sensitive to shunt capacitance at the input.

A resistive part of the source impedance causes additional low-frequency local feedback in the CC stage, driving the poles back towards each other. The pole splitting is effective only if the stage is driven from a very high source impedance.

The local-feedback pole splitting is an effective method for changing the pole positions of the loop transfer function. There is no significant loss of low-frequency loop gain. It should be noted, however, that the high-frequency transimpedance of the described stages is determined by the non-linear and inaccurate capacitance $C_{b'c}$. Although the stage transfers are determined by the negative-feedback action of this capacitance, they are not linearised. When the technique is applied to the output stage, the loss of high-frequency loop gain is not exchanged for transfer quality. When applied to the input stage, the technique does not reduce the SNR, as long as no external capacitances are used and the second-stage contribution can still be ignored.

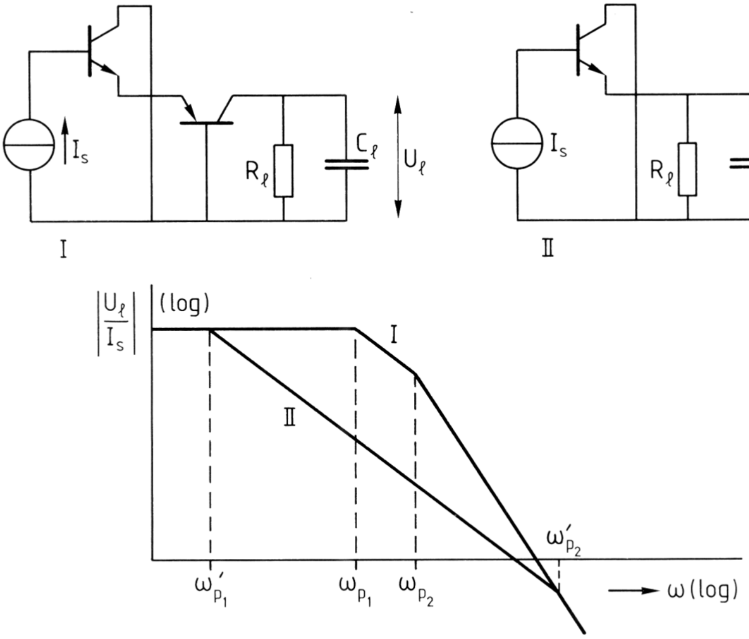


Figure 5.17 Current-driven 'long-tailed pair' with RC load compared with current-driven CC stage with RC load with Bode plot of transimpedance.

5.6.5 Conclusion

In sum, we can conclude that the compensation techniques described thus far offer possibilities of shifting the poles in the loop transfer function. This shift, however, is accompanied by a loss of loop gain somewhere in the frequency region. Depending on where the techniques are applied, non-linearity of the loop gain may be more or less enlarged, or SNR may be deteriorated.

If one has to resort to one of the techniques discussed, shifting of the poles by means of local negative feedback is preferable to shifting by shunt impedances.

In the next sections we will discuss the phantom-zero compensation technique. When feasible, this technique is preferred above the methods of this section.

5.7 Phantom-Zero Techniques

5.7.1 Introduction

In this section we will give some examples of phantom-zero realisations for single-loop amplifier types in connection with sources and loads of various types. We will make no attempt towards completeness, because this would require extensive study most likely resulting in a treatment of active filters.

The approach to phantom-zero realisation probably best begins in an attempt to realise poles in the asymptotic gain $A_{t\infty}$. Unfortunately, these poles do not necessarily give rise to useful phantom zeros because a pole in $A\beta$ may also be created at the location of the phantom zero or in its immediate neighbourhood. In that case the zero is *ineffective*.

Effective phantom-zeros only result if the attenuation of the passive part of the feedback loop (including the source and load impedances) is reduced from the frequency corresponding to the zero and if this reduction is not counteracted immediately by the pole that is simultaneously created. It is frequently possible to realise this situation successfully.

5.7.2 Phantom zeros obtained by series or shunt impedances at the input

(i) Low input impedance

First, we will examine the case where the amplifier has a low input impedance obtained by input shunt feedback as depicted in figure 5.18.

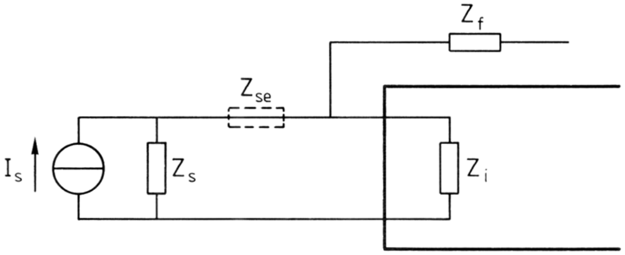


Figure 5.18 Amplifier with input shunt feedback.

- In the case where Z_s is a capacitance C_s, a series resistor R_{se} introduces a pole at $p = -1/R_{se}C_s$ in $A_{t\infty}$ which will be an effective phantom zero, provided that $|Z_s|$ is small with respect to $|Z_i|$. It is possible to realise two complex phantom zeros by using an additional inductance in Z_{se}.
- In the case where Z_s is a resistance, a series inductance L_{se} introduces a pole at $p = -R_s/L_{se}$ in $A_{t\infty}$ A condition similar to the above must apply for the phantom

zero to be effective.

- In the case where Z_s is inductive, there is no possibility of realising a pole in $A_{t\infty}$ with the aid of a series impedance.

(ii) High input impedance

The second case refers to amplifiers with a high input impedance obtained by input series feedback.

- In the case of capacitive sources there is no simple way to realise a pole in $A_{t\infty}$.
- With a resistive source, a shunt capacitance at the input can be used for this purpose. Because the source impedance is frequently rather 'loosely coupled' to the feedback loop in the case of series feedback, the resulting phantom zero will usually not be very effective.
- An inductive source requires an input shunt resistance in order to realise a pole in $A_{t\infty}$. A shunt capacitance may be used to obtain two complex poles. The phantom zero(s) may be not very effective for the same reason as given above.

The series and shunt impedances proposed to be used here at the input for realising poles in the asymptotic gain $A_{t\infty}$ may of course inadmissibly deteriorate the SNR, albeit only by their own noise production. The influence of the amplifier sources is increased a little bit at the band edge, because only there does the impedance as seen from the amplifier become different from the source impedance.

5.7.3 Phantom zeros obtained by series and shunt impedances at the output

(i) Low output impedance

We will first look at the case where a low output impedance is realised with the aid of output shunt feedback. This situation is schematically shown in figure 5.19.

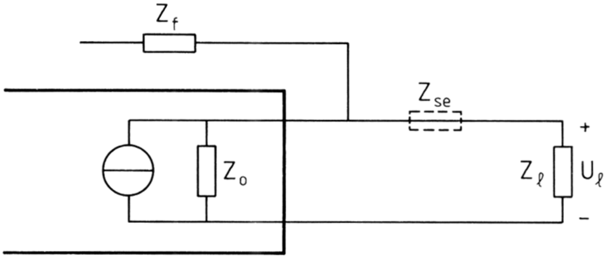


Figure 5.19 Amplifier with output shunt feedback.

- When the load impedance is a capacitance, it is possible to obtain a pole in the asymptotic gain with the aid of an output series resistor R_{se}. Whether this pole causes an effective phantom zero or not depends on the magnitudes of the impedances Z_o, Z_f and Z_ℓ. With the aid of an inductance in Z_{se} complex phantom zeros may be realisable as well.

- A resistive load impedance requires the use of a series inductance for the realisation of a pole in $A_{t\infty}$. The phantom zero is effective under the same condition as given above.
- An inductive load impedance does not admit phantom-zero realisation with a series impedance.

(ii) High output impedance.

Secondly, we will briefly examine the case where a high output impedance is realised with the aid of output series feedback.

- A capacitive load does not admit a phantom zero realisation with a shunt impedance.
- A resistive load requires a shunt capacitance.
- An inductive load requires a shunt resistance.

Because the load impedance in the case of output series feedback is frequently ‘loosely coupled’ to the feedback loop, the phantom zeros may not be very effective. The series and shunt impedances at the output used to obtain phantom zeros may have a slightly adverse effect on power efficiency. Only at the band edge, however, will this become noticeable because there these impedances have the same order of magnitude as the load impedances.

5.7.4 Phantom zeros in the feedback network

Perhaps more obvious possibilities of realising phantom zeros are available in the networks that fix the transfer parameters of the amplifiers which will, for short, be called feedback networks. We will deal exclusively with the impedance-feedback single-loop amplifiers.

(i) Voltage amplifier.

The first configuration to be examined is the voltage amplifier depicted in figure 5.20, where the asymptotic value of the voltage gain U_ℓ/U_s is given by

$$A_{u\infty} = \frac{Z_1 + Z_2}{Z_1}$$

Assuming that a frequency-independent voltage gain has to be realised, we use the same types of impedances for Z_1 and Z_2.

- In the case where Z_1 and Z_2 are resistances, poles in $A_{u\infty}$ can be realised by inserting a capacitance in parallel with R_1 or an inductance in series with R_2. Whether these poles result in effective phantom zeros depends on the ratio of

R_1 and R_2. Both the impedance in parallel with R_1 and the one from which the network is driven play important roles as well.

- When Z_1 and Z_2 are predominantly capacitive a pole in $A_{u\infty}$ can be obtained by inserting a resistor in series with Z_1.
- The case where Z_1 and Z_2 are predominantly inductive seems rather unpractical and will not be dealt with.

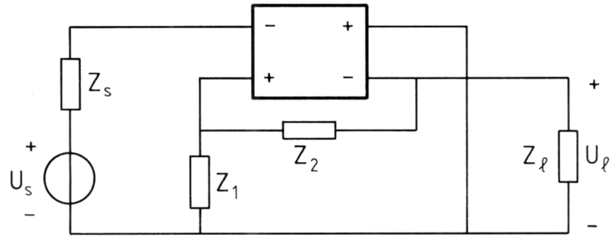


Figure 5.20 Impedance-feedback voltage amplifier.

(ii) Transadmittance amplifier

The transadmittance amplifier depicted in figure 5.21 permits only one element to be influenced for the realisation of a pole in $Y_\infty = -1/Z$. If Y is a capacitance, a resistor can be taken in series. If Y is a resistance, an inductance can be used in series. Whether the phantom zeros thus obtained can be effective depends on the impedance from which Z is driven and by which it is loaded.

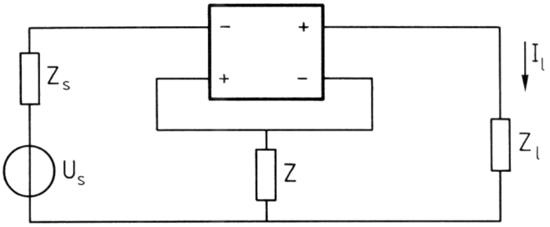


Figure 5.21 Impedance-feedback transadmittance amplifier.

More possibilities for poles in Y_∞ are created by taking a π-attenuator instead of a single impedance. For the same value of the transadmittance we then have to take larger impedances in series with the input and output, which may adversely influence the SNR and PE.

(iii) Transimpedance amplifier

One element can be altered in order to introduce a pole in $Z_\infty = -Z_f$ (figure 5.22) in this case as well. When Z_f is a capacitor, there is no possibility of realising such a pole. In the case where Z_f is a resistor, a shunt capacitance is needed, and in the case of an inductance a shunt resistor. The effectiveness of the phantom zero is determined here as well by the impedance from which Z_f is driven and by which it is loaded. When a T-attenuator is used instead of Z_f, more possibilities for poles in Z_∞ are obtained. The shunt impedances at the input and output are smaller in that case and the SNR and PE may be influenced even more.

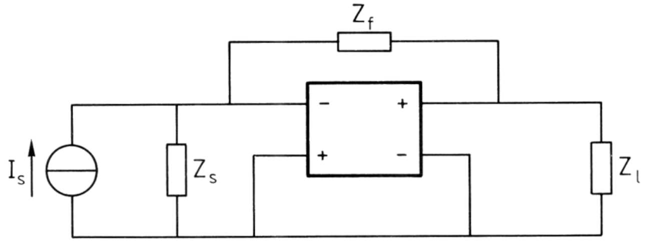


Figure 5.22 Impedance-feedback transimpedance amplifier.

(iv) Current amplifier

The asymptotic value of the current gain of the amplifier of figure 5.23 is given by $A_{i\infty} = I_\ell/I_s = (Z_1 + Z_2)/Z_1$, which is in the same form as the gain of the voltage amplifier (i). The considerations for realising effective phantom zeros are therefore very much similar to those given there.

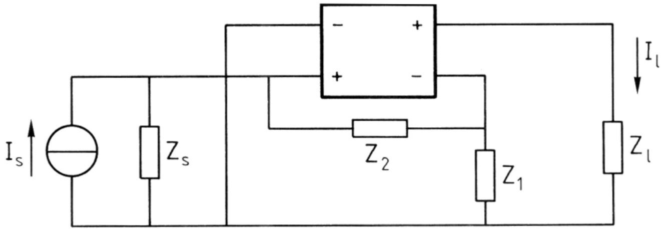


Figure 5.23 Impedance-feedback current amplifier.

5.7.5 Concluding remarks

We have presented in this section some techniques for realising phantom zeros in single-loop negative-feedback amplifiers.

With multiloop amplifiers additional problems arise because not only must the transfer function have an acceptable high-frequency performance but the impedance functions must be well behaved too.

The compensation techniques available for these types of amplifiers are similar to those for single-loop amplifiers. The simultaneous realisation of well-behaved transfer and impedance functions can be based on the theory given in section 1.3.4. Dealing in detail with these design problems, lies beyond the scope of this work.

5.8 Loop Transfer Functions with two Poles

5.8.1 Introduction

We will now discuss in more quantitative terms the design of a transfer function of an amplifier with a loop transfer function $A\beta$ having two poles, which can be denoted as:

$$A\beta = A_o\beta_o \frac{1 - p/n}{(1 - p/p_1)(1 - p/p_2)} \tag{5.7}$$

where $A_o\beta_o$ is the gain at midband frequencies, and n is a phantom zero which may or may not be realisable.
The poles of A_t are the roots of $1 - A\beta = 0$ or:

$$p^2 - p\left(p_1 + p_2 - \frac{p_1p_2}{n} A_o\beta_o\right) + (1 - A_o\beta_o)p_1p_2 = 0 \tag{5.8}$$

The relevant part of the locus of the poles when the L-P product $(1 - A_o\beta_o)p_1p_2 = \omega_n^2$ is a constant is a semi-circle with radius ω_n together with the negative real axis as depicted in figure 5.24. We will refer to it as the *product locus*.

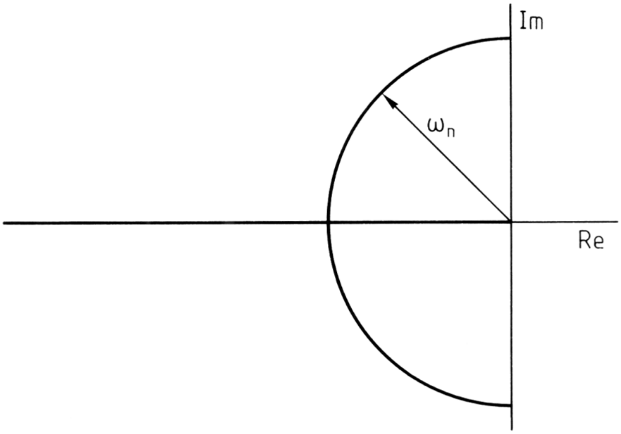


Figure 5.24 Locus of the poles when ω_n is a constant ('product locus').

The positions of the poles on this locus depend on the coefficient of p in (5.8). In order to force the poles of A_t into the desired Butterworth positions the following condition must be met:

$$p_1 + p_2 - \frac{p_1p_2}{n} \cdot A_o\beta_o \approx p_1 + p_2 + \frac{\omega_n^2}{n} = -\sqrt{2} \cdot \omega_n \tag{5.9}$$

where

$$\omega_n^2 = (1 - A_o\beta_o)p_1p_2 \approx -A_o\beta_o p_1 p_2$$

If it is not feasible to realise a phantom zero, the sum of the poles must have a value

$$p_1 + p_2 = -\sqrt{2} \cdot \omega_n^2$$

and to accomplish this the positions of either one or both of the poles will generally have to be changed, preferably without changing the loop-gain-poles product. One of the compensation techniques of the previous sections may be used for this purpose.

Figure 5.25 gives an example. The root locus intersects the 'product locus' in undesired positions, marked by ⊠ . The locations of the poles of AD must be changed so that the root locus intersects the product locus in the desired positions, marked by ⊗ .

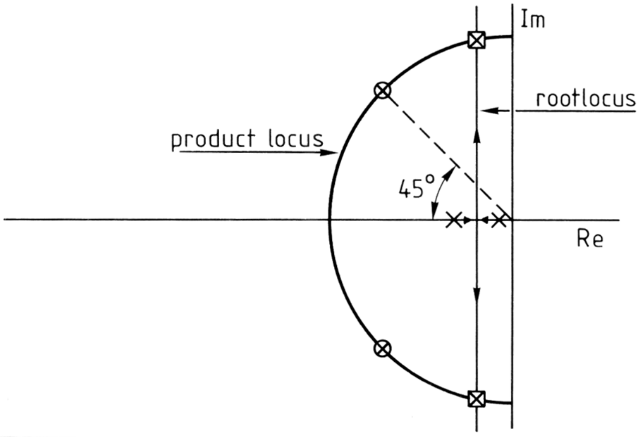


Figure 5.25 Root locus and 'product locus' of a two-poles system.
⊠ actual undesired pole positions
⊗ desired Butterworth positions.

5.8.2 Compensation with a phantom zero

In section 5.2.2 we saw that for the compensation of a two-poles loop transfer function a phantom zero n can be used to move the poles of A_t into Butterworth positions. The required value of this zero follows from (5.9) and is given by:

$$n \approx \frac{-\omega_n^2}{p_1 + p_2 + \omega_n\sqrt{2}}$$

If it is feasible to realise an effective phantom zero, the positions of the poles of the loop transfer function are arbitrary, provided that:

$$|p_1 + p_2| < \sqrt{2} \cdot \omega_n$$

As we have seen in section 5.7, adding a phantom zero means that the attenuation of the passive part of the feedback loop is reduced beyond the frequency corresponding to this zero. Hence, the loop gain is enlarged unlike in the other compensation techniques, and therefore, it is considered the best possible method for shaping the root locus in the desired way. The increase in the loop gain resulting from a phantom-zero compensation is illustrated in the Bode plot of figure 5.26. Just before the loop gain is reduced to unity the frequency roll-off is reduced to a 6dB-per-octave slope again.

A reduction in the L-P product cannot always be avoided when adding a phantom zero. It goes without saying that this reduction should be as small as possible.

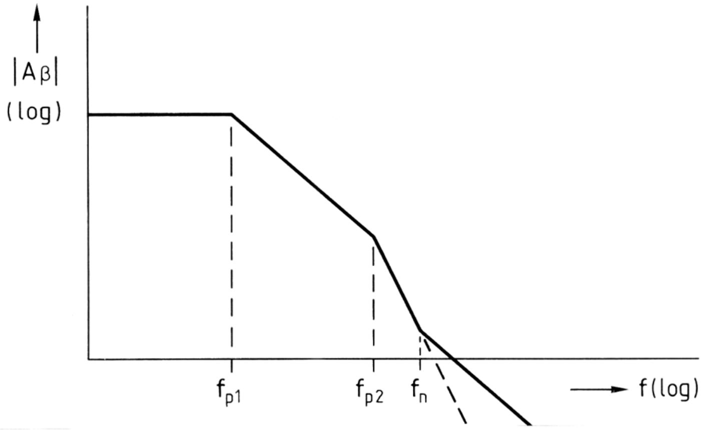


Figure 5.26 Bode plot of a phantom-zero compensated loop transfer function with two poles.

5.8.3 Influence of an additional pole on the transfer function of an amplifier, designed as a second-order system

Technical systems will frequently have additional poles that do not belong to the dominant group. Moreover, it may happen that the sum of the poles is too large for obtaining an MFM response ($|p_1 + p_2| > \sqrt{2\omega_n}$) In that case the loop gain can be increased and a stage that contributes a non-dominant pole may be added.

We did not yet give criteria for judging whether or not a pole belongs to the dominant group. In this section we will examine the influence of one additional pole in a technical second-order system in order to find these criteria.

The loop transfer function of such an amplifier can be written as:

$$A\beta = A_o\beta_o \frac{1 - p/n}{(1 - p/p_1)(1 - p/p_2)} \cdot \frac{1}{1 - p/p_3}$$

where n is supposed to be a phantom zero.

When p_3 is disregarded, the transfer function A_t is designed as an MFM function with:

$$\omega_n^2 = (1 - A_o\beta_o)p_1p_2 \qquad n = \frac{-\omega_n^2}{p_1 + p_2 + \omega_n\sqrt{2}}$$

The denominator of A_t can then be written as:

$$D_{2A_t} = 1 + \sqrt{2}\frac{p}{\omega_n} + \frac{p^2}{\omega_n^2}$$

The modulus of the denominator determines the course of the frequency response. For the MFM function with two poles it is given by:

$$|D_{2A_t}|^2 = \left(1 - \frac{\omega^2}{\omega_n^2}\right)^2 + 2\frac{\omega^2}{\omega_n^2}$$

The general form of the denominator in the case of three poles is:

$$D_{A_t} = 1 + pa_{p_1} + p^2a_{p_2} + p^3a_{p_3}$$

and $$|D_{A_t}|^2 = \{1 - \omega^2a_{p_1}\}^2 + \omega^2\{a_{p_1} - \omega^2a_{p_3}\}^2$$

An increase in the factors a_{p_2} and a_{p_3} obviously enlarges the transfer in a certain frequency region and may result in peaking of the transfer characteristics.

Considering now the influence of a third pole p_3, in a transfer function designed as a second-order MFM-transfer, we can write the denominator of A_t as:

$$D_{A_t} = 1 - p\left\{\frac{p_1p_2 + p_1p_3 + p_2p_3}{\omega_n^2p_3} + \frac{1}{n}\right\} + p^2\left\{\frac{1}{\omega_n^2}\left(1 + \frac{p_1 + p_2}{p_3}\right)\right\} - p^3\frac{1}{p_3\omega_n^2}$$

Because p_3 was assumed to be non-dominant ($|p_3| \gg |p_1|, |p_2|$), it follows that:

$$D_{A_t} = D_{2A_t} + p^2\left\{\frac{\sqrt{2}}{\omega_n} - \frac{1}{n}\right\}\frac{1}{p_3} - \frac{p^3}{p_3\omega_n^2}$$

The influence of p_3 on the frequency response will be examined for two situations. In the first case, we assume that $p_1 \approx p_2 < \omega_n$ so that the phantom zero n has a value $n = -\omega_n/\sqrt{2}$. In the second case, the poles p_1 and p_2 have values such that no phantom zero is needed. In the case where the phantom zero is needed, we find:

$$D_{A_t} = D_{2A_t} - \frac{p^3}{p_3\omega_n^2}$$

and
$$|D_{A_t}|^2 = \left(1 - \frac{\omega^2}{\omega_n^2}\right)^2 + \omega^2\left(\frac{\sqrt{2}}{\omega_n} + \frac{\omega^2}{\omega_n^2}\frac{1}{p_3}\right)^2 \tag{5.10}$$

A reduction in the denominator is observed because the factor a_{p_3} is increased, resulting — as might be expected — in a peaking of the frequency response.
In the case where no phantom zero is used a stronger influence of the additional pole is found, as appears from the expression of the modulus of the denominator:

$$|D_{A_t}|^2 = \left\{1 - \omega^2\left(\frac{1}{\omega_n^2} - \frac{\sqrt{2}}{p_3}\frac{1}{\omega_n}\right)\right\}^2 + \omega^2\left(\frac{\sqrt{2}}{\omega_n} + \frac{\omega^2}{\omega_n^2}\frac{1}{p_3}\right)^2 \tag{5.11}$$

The frequency responses were calculated for normalised values of ω_n, and are given in figures 5.27a and 5.27b for the cases described by (5.10) and (5.11), respectively. A significantly lower sensitivity for the presence of the third pole as far as the peaking is concerned is found for the case where a phantom zero is used (figure 5.27).
Figures 5.28a and 5.28b show the influence of a 25% variation in the L-P product. The system with the phantom zero is clearly less sensitive to variations in the circuit parameters.

5.8.4 Conclusions

We have demonstrated in this section that, in the case of a loop transfer function with two dominant poles, the poles of the amplifier transfer function can be manoeuvred into the desired Butterworth positions either by giving the $A\beta$ poles such values that the root locus intersects the product locus at the Butterworth locations, or by using a phantom zero that shapes the root locus in the desired way.
When for a given poles product the sum of the poles of $A\beta$ is too small to realise the required Butterworth positions, it is possible to enlarge the loop gain with the aid of a stage that produces a non-dominant pole. This pole is allowed to have a magnitude of about five times ω_n. In other words, an excess phase of about 10 degrees can be excepted, also in the case where more than one non-dominant pole is involved.
It was shown that the phantom-zero compensation technique is preferable to other techniques for two reasons:

1. The loop gain is increased instead of reduced.
2. A lower sensitivity to parameter variations in the amplifier frequency response is obtained.

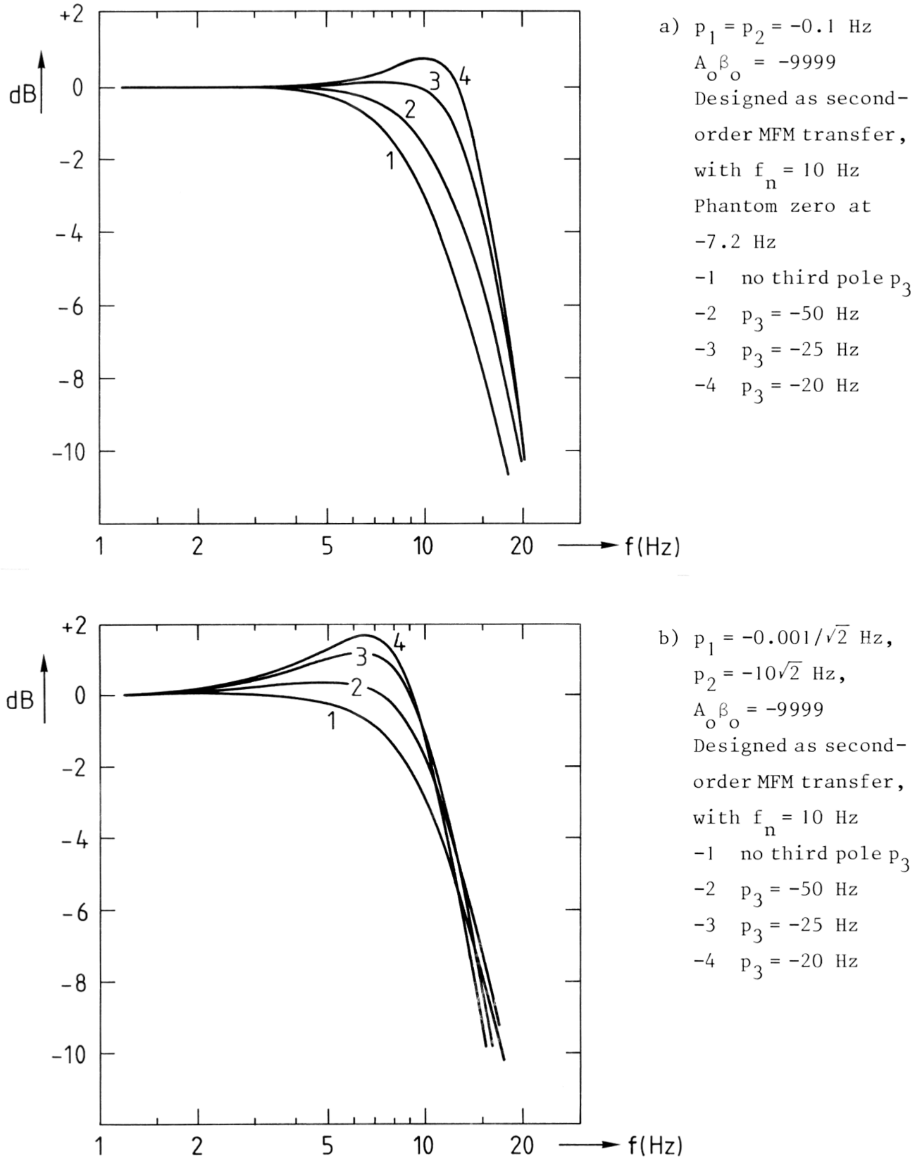
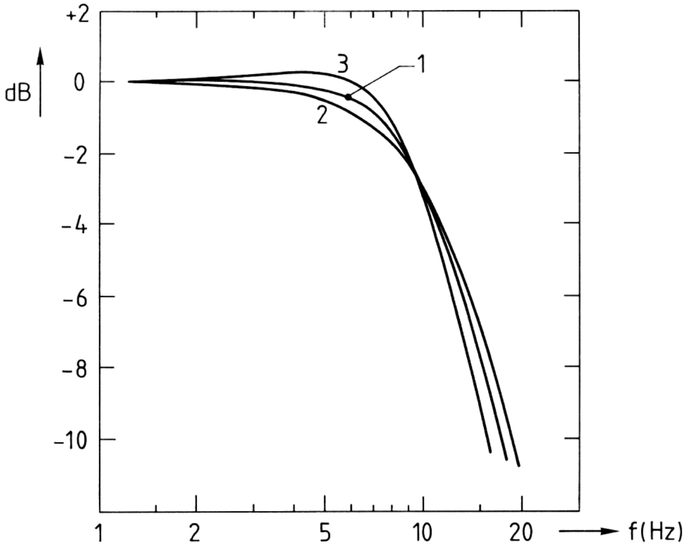
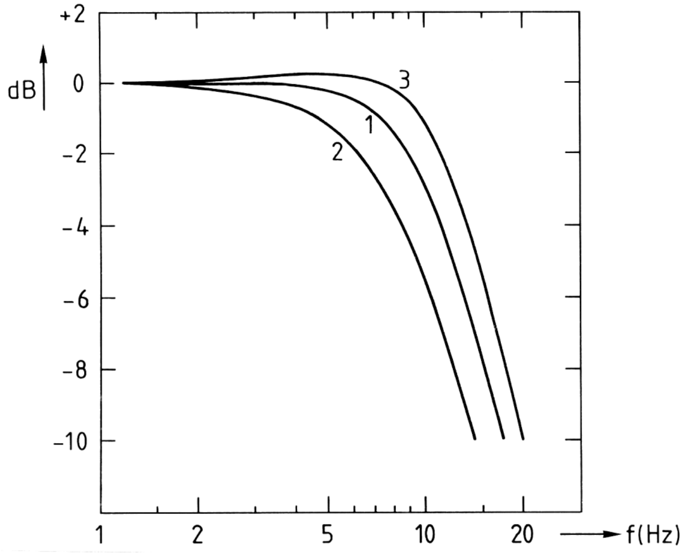


Figure 5.27 *Influence of a third pole on the frequency response of a system designed as an MFM two-poles response.*
a) $p_1 = p_2 = -0.1$ Hz, $A_o\beta_o = -9999$. Designed as second-order MFM transfer, with $f_n = 10$ Hz. Phantom zero at –7.2 Hz. 1. no third pole p_3; 2. $p_3 = -50$ Hz; 3. $p_3 = -25$ Hz; 4. $p_3 = -20$ Hz.
b) $p_1 = -0.001/\sqrt{2}$ Hz, $p_2 = -10\sqrt{2}$ Hz, $A_o\beta_o = -9999$. Designed as second-order MFM transfer, with $f_n = 10$ Hz. 1. no third pole p_3; 2. $p_3 = -50$ Hz; 3. $p_3 = -25$ Hz; 4. $p_3 = -20$ Hz.



a. With phantom zero.



b. Without phantom zero.

Figure 5.28 *Influence of a ±25% variation in the L-P product on the transfer function of a second-order system.*
a) With phantom zero; b) without phantom zero.

5.9 Loop Transfer Functions with three Poles

We will now continue our treatment with a discussion of the design of an amplifier transfer function with three poles. An all-pole MFM response will be our design goal again. The third-order Butterworth polynomial can be written as:

$$p^3 + 2\omega_n p^2 + 2\omega_n^2 + \omega_n^3 = 0$$

We will first discuss the case where $A\beta$ does not contain any zeros. Making the coefficients in the equation $1 - A\beta = 0$ equal to the coefficients in this polynomial yields:

$$(1 - A_o\beta_o)p_1p_2p_3 = -\omega_n^3 \tag{5.12}$$

$$p_1p_2 + p_1p_3 + p_2p_3 = 2\omega_n^2 \tag{5.13}$$

$$p_1 + p_2 + p_3 = -2\omega_n^2 \tag{5.14}$$

From equations (5.13) and (5.14) it follows that

$$p_1^2 + p_2^2 + p_3^2 = 0$$

which can only be true if two roots are complex conjugates.
With $p_2 = \alpha + j\beta$ and $p_3 = \alpha - j\beta$, it follows that β must be larger than or equal to α:

$$\beta \geq \alpha \tag{5.15}$$

From equation (5.14) it follows that for $p_1 < 0$, the condition must be met that:

$$-\omega_n \leq \alpha \leq 0 \tag{5.16}$$

A third condition is found from equations (5.12), (5.13) and (5.14):

$$(\omega_n + \alpha)\{\omega_n^2 + 2\alpha(\omega_n + \alpha)\} = \frac{\omega_n^3}{4A_o\beta_o} \tag{5.17}$$

With condition (5.16), condition (5.17) can only be met if $\alpha = -\omega_n$ and $A_o\beta_o \to -\infty$. In that case, p_1 must equal zero and β must equal $-\omega_n$.
Figure 5.29 shows the positions of the poles of the loop transfer function required for realising a Butterworth characteristic and the corresponding root locus.

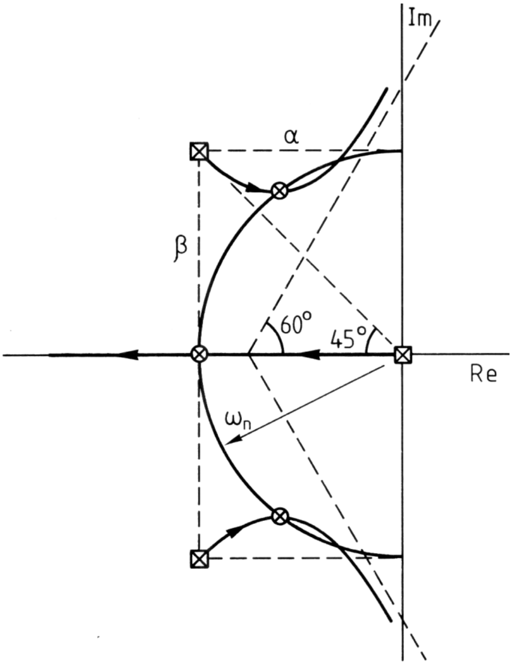


Figure 5.29 Pole positions ☒ and root locus of the loop transfer required for realisation of an MFM characteristic when no phantom zeros are available.

Giving an example of a possible structure of the active part of an amplifier may be clarifying. A schematic circuit representing a so-called charge amplifier with capacitive source and load which might be designed as described above is given in figure 5.30.

A pole in the origin and an infinite low-frequency loop gain are approximated here, by a current-driven CS stage. Some degradation in its contribution to the L-P product (being ideally equal to $\omega_T = g_m/(C_{gs} + C_{gd})$) can hardly be avoided because of parasitic shunt capacitance. The complex pole pair is realised by using a negative-feedback stage, for instance a shunt stage.

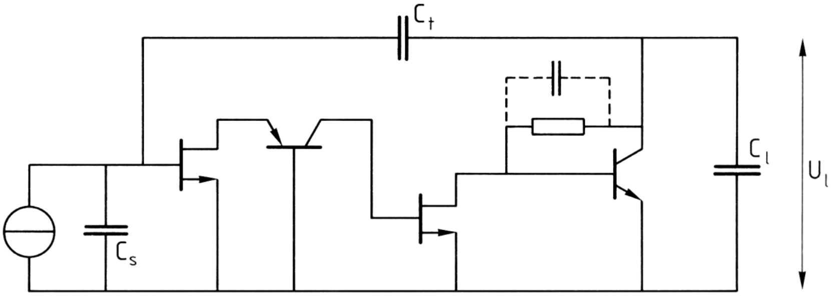


Figure 5.30 Charge amplifier with capacitive load. The loop transfer function can have the singularity pattern as given in figure 5.28.

It should be noted that the amplifier type of this example usually offers other more preferable possibilities for obtaining an MFM characteristic by using phantom zeros.

In many situations, a phantom zero can be added to the loop transfer function, without seriously disturbing the positions of the poles. The following conditions must be met for the realisation of an MFM characteristic in the case of one available phantom zero n, which offers an additional degree of freedom.

$$(1 - A_o\beta_o)p_1p_2p_3 = -\omega_n^3 \approx -A_o\beta_o p_1p_2p_3 \tag{5.18}$$

$$p_1p_2 + p_1p_3 + p_2p_3 - \omega_n^3/n = 2\omega_n^2 \tag{5.19}$$

$$p_1 + p_2 + p_3 = -2\omega_n \tag{5.20}$$

From (5.20) it is seen that the positions of the poles of $A\beta$ cannot be chosen arbitrarily. Compensation techniques as given in section 5.6 will generally be applicable for meeting this condition. The zero can be chosen so that condition (5.19) is met and the flexibility of the design is increased.

A still larger increase in flexibility is obtained if two effective phantom zeros can be added to the loop transfer function. The MFM conditions then become:

$$(1 - A_o\beta_o)p_1p_2p_3 = -\omega_n^3 \approx -A_o\beta_o p_1p_2p_3$$

$$p_1p_2 + p_1p_3 + p_2p_3 + \omega_n^3(1/n_1 + 1/n_2) = 2\omega_n^2$$

$$p_1 + p_2 + p_3 - \omega_n^3/n_1n_2 = -2\omega_n$$

The conditions for the feasibility of realisation are now:

$$|\Sigma p| \leq 2\omega_n$$

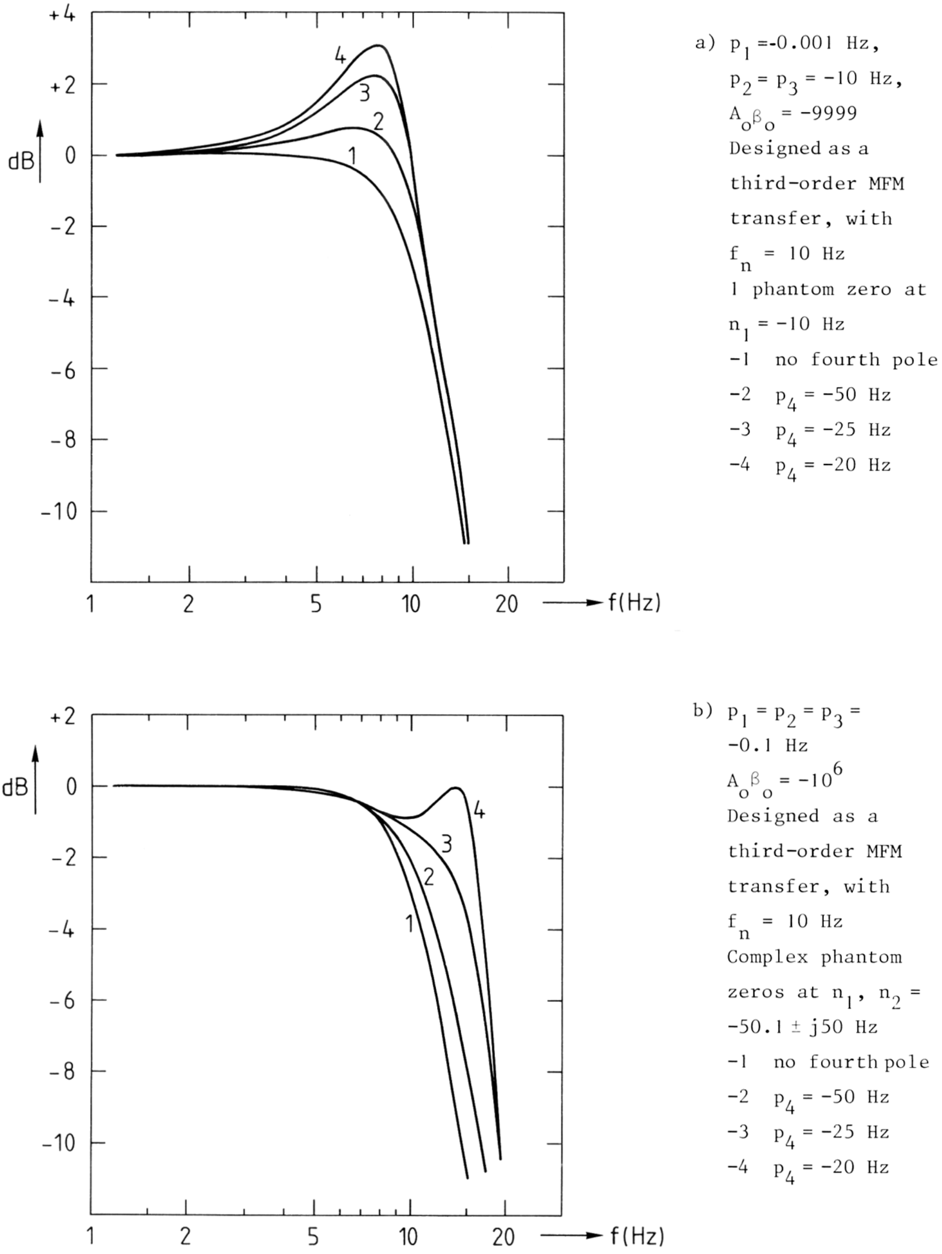
$$p_1p_2 + p_1p_3 + p_2p_3 < 2\omega_n^2$$

Sometimes MFM realisation is only possible with complex phantom zeros.

Examples of frequency responses of amplifiers that are designed as third-order systems are given in figure 5.31, where ω_n is normalised at $2\pi\cdot 10$ rad/sec.

Responses are also shown with an additional fourth pole with a magnitude of 2, 2.5 and 5 times the normalised bandwidth.

It also appears here that a pole with a magnitude of about five times the value of ω_n (calculated for three poles in the dominant group) can reasonably be considered as non-dominant.



a) p_1 = -0.001 Hz,
$p_2 = p_3$ = -10 Hz,
$A_o\beta_o$ = -9999
Designed as a
third-order MFM
transfer, with
f_n = 10 Hz
1 phantom zero at
n_1 = -10 Hz
-1 no fourth pole
-2 p_4 = -50 Hz
-3 p_4 = -25 Hz
-4 p_4 = -20 Hz

b) $p_1 = p_2 = p_3$ =
-0.1 Hz
$A_o\beta_o = -10^6$
Designed as a
third-order MFM
transfer, with
f_n = 10 Hz
Complex phantom
zeros at n_1, n_2 =
-50.1 ± j50 Hz
-1 no fourth pole
-2 p_4 = -50 Hz
-3 p_4 = -25 Hz
-4 p_4 = -20 Hz

Figure 5.31 *Influence of a fourth pole on the frequency response of a system designed as an MFM three-poles response.*
a) $p_1 = -0.001$ Hz, $p_2 = p_3 = -10$ Hz, $A_o\boldsymbol{\beta}_o = -9999$. Designed as third-order MFM transfer, with $f_n = 10$ Hz. 1 Phantom zero at $n_1 = -10$ Hz. 1. no fourth pole; 2. $p_4 = -50$ Hz; 3. $p_4 = -25$ Hz; 4. $p_4 = -20$ Hz.
b) $p_1 = p_2 = p_3 = -0.1$ Hz, $A_o\boldsymbol{\beta}_o = -10^6$. Designed as third-order MFM transfer, with $f_n = 10$ Hz. Complex phantom zeros at n_1, $n_2 = -50.1 \pm j50$ Hz. 1. no fourth pole; 2. $p_4 = -50$ Hz; 3. $p_4 = -25$ Hz; 4. $p_4 = -20$ Hz.

5.10 Discussion

In the previous sections we presented design considerations and techniques for the optimum realisation of transfer functions of negative-feedback amplifiers with respect to their high-frequency performance. The all-pole MFM approximation for the transfer-function filter characteristics was chosen as a vehicle, because its parameters are related to the bandwidth in a simple way.

It was found with the aid of the asymptotic-gain model that the product of the loop gain and the magnitudes of the dominant poles (L-P product) is a useful measure for the realisable bandwidth. The contribution a stage can give to this product is therefore a figure of merit for that stage. Characterisation of this contribution with the classical gain-bandwidth product is considered to be misleading, the reason why this characterisation of bandwidth capabilities is abandoned in this work.

Well-designed stages can give a contribution ω_T to the L-P product. Criteria for a well-designed stage appear to be that no small shunt impedances or large series impedances should be present in the signal path. All single-device configurations are able to give approximately the same contributions when these criteria are met. The pole-zero pattern of the loop transfer function is, as a matter of course, strongly dependent on the types of configurations used.

The designability of a feedback loop is considered to be an important aspect. Good designability requires that the loop transfer function be determined by the product of single transfer parameters of stages. This might be accomplished by using small shunt impedances or large series impedances in the signal path, however, with injurious consequences for various quality aspects. Much better results are obtained by using local-feedback stages in such a way that their transfer functions hardly interact. Series stages are preferably avoided in the active part, because they exhibit a relatively bad high-frequency performance.

Combinations of CE and CC stages with CB configurations, resulting in the cascode and the long-tailed pair, respectively, appear to be well behaved and are virtually unilateral for practical source and load conditions. They can frequently be considered as one stage, whose transfer function is characterised by one effective pole, and their contribution to the L-P product amounts to ω_T.

Methods for forcing the root locus through the desired MFM pole positions were discussed. Compensation techniques that use either shunt impedances in the signal path or local feedback were described. Local-feedback compensation techniques are characterised by the realisation of a well-controlled interaction between the input and output circuits of a stage. These are superior to techniques that do not use local feedback.

A very powerful compensation technique is offered by the use of phantom zeros. It is superior to the other techniques because it is not accompanied by a loss of loop gain. The phantom-zero compensation technique excels by offering the possibility of avoiding conflicting requirements on stages with respect to the various quality aspects, because the positions of the poles of the loop transfer function can to a large extent by chosen freely. Consequently, the stages can be optimised with respect to noise, distortion and accuracy performance. The frequency response can be set independently by properly locating the phantom zero(s). Moreover, the transfer function becomes less sensitive to excess phase caused by non-dominant poles and to parameter variations. Unfortunately, the technique cannot be applied in every situation.

References

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6 The Design of Bias Circuitry

6.1 Introduction

In the preceding chapters it was assumed that the bias circuit of an amplifier can be designed in such a way that it has no appreciable influence on the performance of the signal circuit. This assumption implies that the elements of the bias circuit do not change impedance levels in the signal path, do not contribute noise but do prevent the penetration of signals to and from the power supply leads. A systematic approach to the design of such bias circuitry is, to our knowledge, not available.

General reflections on bias-circuit design are almost exclusively found in textbooks. These design considerations frequently concern single active devices and are mainly devoted to the most basic function of the bias circuits, namely maintaining the intended DC currents and voltages.

The types of circuits that generally emerge have small shunt resistances at input and output. Figure 6.1 shows the most wide-spread example of a bipolar-transistor bias circuit, where four resistances and a supply voltage source U_B are used to establish the currents and voltages.

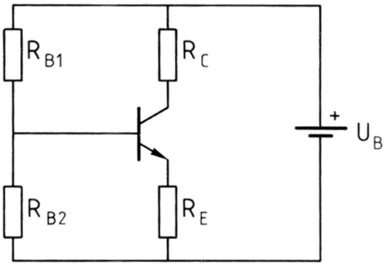


Figure 6.1 Simple biasing circuit for a bipolar transistor.

For stable biasing, the voltage across R_E should be large, whereas the shunt resistance formed by R_{B1} and R_{B2} should be small. The collector-to-emitter voltage must be positive in this case, and therefore, roughly the following condition must be met:

$$U_{CE} \approx U_B - I_E(R_E + R_C) \geq 0$$

assuming that I_C and I_E are approximately equal.

The values of R_C — the so-called load resistance in a CE stage — and of R_E are obviously limited by the values of the emitter current and the supply voltage. Little freedom is left for the design of a desired signal transfer, unless large inductances and transformers in addition to bypass capacitors are used. In many situations, inductive elements are not appropriate for use in wide-band amplifiers. Consequently, available signal power is lost in the shunt resistances at input and output and thermal noise is generated.

It is an understatement to say that the bias circuit of figure 6.1 is not ideal from the viewpoint of signal transfer.

A similar bias circuit is usually proposed for an FET; R_{B1} and R_{B2} can, however, be very large in that case.

In this chapter we will briefly discuss a design approach that leads to a more general formulation of bias-circuit design than the usual. This approach departs from the view that the signal path should not be influenced by the bias circuits. The biasing techniques that will be proposed are not essentially new. The explicit formulation of the design method, however, is believed to be unique.

We will restrict ourselves to bias-circuit design for amplifiers that do not process very low-frequency signals (commonly referred to as DC amplifiers). In the amplifiers to be considered here, the information-representing signals can be separated from relatively slow variations in the bias quantities by low-pass or high-pass filters.

It is beyond the scope of this work to develop methods for — in all respects — optimum bias-circuit design, because the character of bias circuitry is determined to a large extent by available or current technology.

6.2 General Design Considerations

For the purpose of biasing a bipolar transistor in the active working region, the relation between currents and voltages can simply but adequately be described by two equations:

$$I_E = I_s \exp \frac{qU_{BE}}{kT}$$

and $$I_C = h_{FE} I_B$$

Though I_E and U_{BE} are functions of U_{CB} or U_{CE} as well, these influences can be regarded as second-order effects in most situations.

In order to achieve biasing in the active region, one of the quantities U_{BE}, I_B or I_E must be supplied to the transistor, while either U_{CE} or U_{CB} must be fixed. If either U_{BE} or I_B is supplied to the transistor, the resulting value of I_E is not accurately known, as a consequence of production tolerances and temperature sensitivity.

Because a number of significant transistor parameters are known as functions of I_E, and because I_E determines the current-handling capability, its required value follows from signal-path design, and it is desirable to fix it reasonably well. Furthermore, U_{CB} — rather than U_{CE} — is considered to have a desired value following from signal-circuit design as well, because the voltage-handling capability is limited primarily by that value of U_{CB} for which the collector-base junction becomes forward biased. However, this distinction is meaningful only when the devices operate with very small U_{CB} values.

Similar considerations are valid for FET's. Drain current and drain-to-gate voltage are the quantities that primarily determine the signal behaviour of the device.

The bias circuit has to be designed such that no shunt and series impedances are needed in the signal path other than those intended for optimum information transfer. These impedances will generally not have the proper values for serving as bias components as well. For the present, we will take the view, therefore, that (controlled) *current sources* and *voltage sources* are allowed exclusively. The implementation of these sources will be briefly discussed in one of the following sections.

First, we will present in the next section a formal approach to the design of the bias circuit of a single transistor that may be incorporated in the active part of an amplifier.

6.3 A Formal Approach to the Biasing of Single Active Devices

6.3.1 Introduction

Each active device in an amplifier needs its own bias circuit, in order to fix its currents and voltages. In the types of amplifiers considered in this work, it is possible to bias each stage individually. A more efficient approach, however, is to let the other amplifier stages be a part of the bias circuit of one particular stage. Both of the above possibilities are left open in the following.

6.3.2 Bias circuit for the CE stage

A biasing scheme for a CE stage, which satisfies the requirement that the bias components have no influence on the signal path, is depicted in figure 6.2. This circuit obviously needs additional means for proper functioning, because of the inaccurately known relation between I_B and I_C in the active region, and because the voltage U_{CE} has virtually no influence on I_C, so that the value of U_{CE} is not well defined.

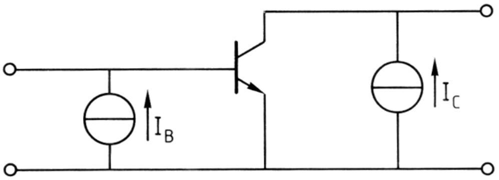


Figure 6.2 DC current sources for biasing a CE stage with an NPN transistor.

In order to fix the value of U_{CB}, the current I_B must be controlled to the proper value as prescribed by the transistor, i.e. $I_B = I_C/h_{FE}$. The value of U_{CE} can be fixed as well instead of fixing U_{CB}, which will yield satisfactory results in the majority of cases, although possibly not when the transistor has to operate with very low bias voltages.

A schematic feedback circuit that accomplishes the desired function is shown in figure 6.3. Load and source conditions are supposed to have no influence on bias conditions. This can be achieved by using coupling capacitors.

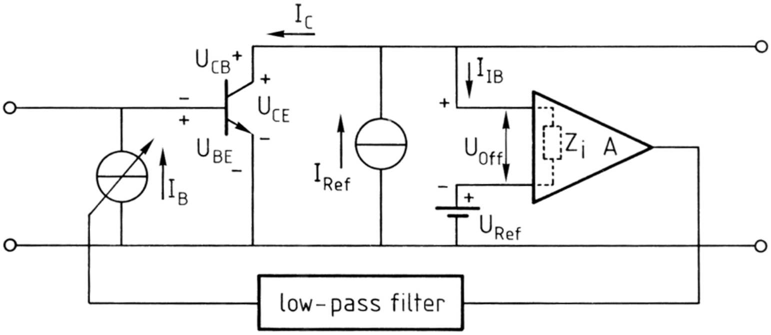


Figure 6.3 Schematic bias circuit for a CE stage.

The collector-to-emitter voltage of the transistor is sensed by the input port of the high-input-impedance amplifier T and compared with the reference voltage U_{Ref}. The amplifier generates the error signal which controls the base bias current I_B to the required value. The amplifier T can be characterised in order to find the quiescent conditions by an input bias current I_{IB} and an input offset voltage U_{Off}. Moreover, it is supposed to have an infinite input impedance Z_i in order not to load the stage. Furthermore, it should be able to handle the signal voltage occurring at the output without introducing non-linearities.

The current source I_B is controlled by the low-pass-filtered output quantity (current or voltage) of the amplifier. The function of the low-pass filter is to avoid a disturbance in the impedance levels at the input and output in the frequency range of the signal information. The operating point of the CE stage is thus fixed at:

$$I_C = I_{Ref} - I_B \quad U_{CB} = U_{Ref} + U_{Off} - U_{BE}$$

For the design of the low-pass filter, the dynamic performance of the bias circuit must be taken into account. It depends, of course, on external conditions, such as the source and load and of the dynamic properties of the transistor and the amplifier T. Requirements for the dynamic performance follow from the desired low-frequency signal performance. We will deal with this design aspect in a later section.

6.3.3 Bias circuits for local-feedback stages

The local-feedback single-device configurations can be biased in a way similar to that of the CE stage. Figure 6.4 shows schematic bias circuits for the single-loop configurations. The resulting bias conditions are given in the figure legends. The current I_B and the voltage U_{Off} are not necessarily small. For accurate biasing, however, either small or accurate values are required.

For each active device in an amplifier, a bias circuit with such a structure must be demonstrable in order to realise proper functioning. The amplifiers T will frequently comprise one or more or even all active devices of the amplifier signal path. In those cases, their input impedances are not necessarily high. The controlled current source may be incorporated in the configuration intended to drive the stage which is to be biased. In that case, it belongs to the signal path as well. Of course, the amplifier T itself must be biased too.

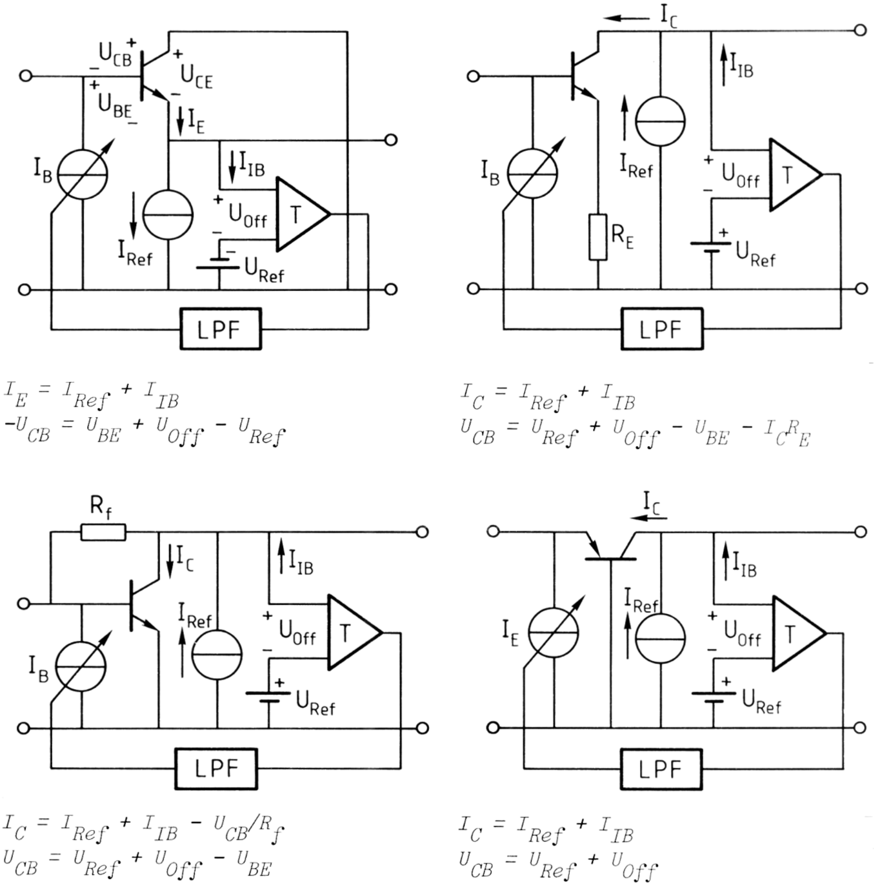
The reference currents and voltages that are needed for proper biasing are generally derived from one or two supply voltages (supply currents would be equally possible [1], but are very unusual). The availability of two supply voltages simplifies bias-circuit design in many situations.

The bias-feedback circuits shown in figures 6.3 and 6.4 are all of the same types. The output voltage is sensed and the input current is controlled. A *transimpedance* type of transfer function therefore results in all cases in the frequency region where the low-pass filter does not prevent the feedback action.

6.4 Bias Circuit for a Complete Amplifier

6.4.1 Introduction

In order to bias a complete amplifier, it is *possible* though perhaps not always practical to use an overall feedback loop that has a function similar to that of the loops for single active devices. In order to fix the output bias voltage for guaranteeing adequate signal-handling capability, the input bias current must be controlled by a current or voltage derived directly from the output voltage. Figure 6.5 shows the resulting general configuration, where the ABCD block refers to the negative-feedback amplifier of which one or more transmission parameters are fixed.



$I_E = I_{Ref} + I_{IB}$

$-U_{CB} = U_{BE} + U_{Off} - U_{Ref}$

$I_C = I_{Ref} + I_{IB}$

$U_{CB} = U_{Ref} + U_{Off} - U_{BE} - I_C R_E$

$I_C = I_{Ref} + I_{IB} - U_{CB}/R_f$

$U_{CB} = U_{Ref} + U_{Off} - U_{BE}$

$I_C = I_{Ref} + I_{IB}$

$U_{CB} = U_{Ref} + U_{Off}$

Figure 6.4 Bias circuits for single loop feedback configurations.

Each active device in the amplifier is supposed to have a bias circuit with a structure as given in figures 6.3 and 6.4. The amplifiers T, needed *there* to bias a particular device, are supposed to comprise *all other* amplifier stages, in addition to the external bias loop. Only the amplifier T of the external loop needs to have an infinite input impedance in this case.

Capacitors are used in figure 6.5 to connect the source and the load, in order that the bias quantities not be disturbed and that no direct currents flow through the source and load. Though it will not always be necessary to disconnect the source and load from the DC-bias circuit, we will discuss only this situation here.

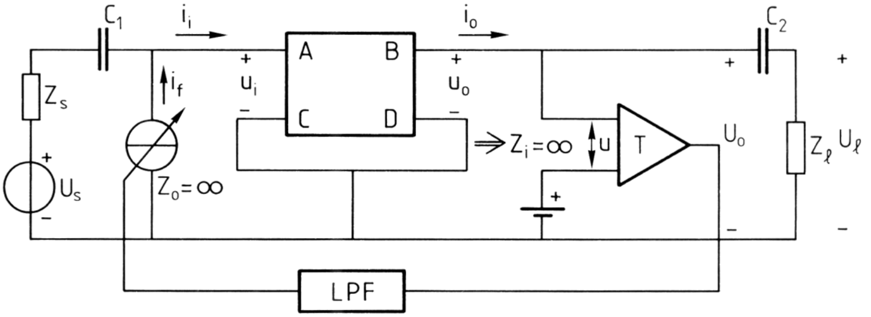


Figure 6.5 General form of 'external' bias circuit.

6.4.2 Modeling the influence of the basic bias circuit on signal performance

For finding a model that adequately describes the influence of the bias circuit, the signal performance at all but low frequencies is assumed to be determined by the amplifier block characterised by its transmission parameters A, B, C and D (figure 6.5). One or more of these parameters may be fixed by the signal-feedback loop. The transfer function U_ℓ/U_s is considered as an example for setting up the model. The voltage transfer function U_o/U_s, when determined entirely by the signal path, has the form (1.10):

$$\frac{U_o}{U_s} = A_u = \frac{Z'_\ell}{AZ'_\ell + B + CZ'_sZ'_\ell + DZ'_s} \tag{6.1}$$

where Z' and Z'_s are the load and source impedances, including the impedances of the blocking capacitors C_2 and C_1, respectively.
The bias feedback loop changes this transfer function into:

$$\frac{U_o}{U_s} = A'_u = \frac{Z'_\ell}{AZ'_\ell + B + (C+g)Z'_sZ'_\ell + DZ'_s} \tag{6.2}$$

where g is a (low-pass) transfer function defined as:

$$g = \frac{i_f}{u} \quad \text{(See figure 6.5)}$$

Expression (6.2) can alternatively be written as:

$$\frac{U_o}{U_s} = A'_u = \frac{A_u}{1 + A_u g Z'_s} = \frac{A_u}{1 + A_u k}$$

The transfer function U_ℓ/U_s is found by multiplying A'_u by the factor $pC_2Z_\ell/(1 + pC_2Z_\ell)$, so that

$$\frac{U_\ell}{U_s} = \frac{A_u}{1 + A_u k}\left(\frac{pC_2 Z_\ell}{1 + pC_2 Z_\ell}\right)$$

Other transfer functions have similar forms.

It appears to be possible to describe the influence of the bias circuit by using Black's elementary feedback model, which was discussed in chapter 1. The dynamic performance of the bias loop will be discussed in the next section on the basis of this feedback model.

It is useful to note that it can be seen from expression (6.2) that an amplifier with a feedback-determined transimpedance parameter $1/C$ may not need the external bias-feedback loop with its transfer function g, because g plays a role similar to the reciprocal transimpedance C. When the feedback loop that determines C has a DC path, the g-loop may be superfluous. In the case of an integrating transimpedance or charge amplifier, a large resistance can be used in parallel with the feedback capacitance [2].

6.4.3 Dynamic performance of the biased amplifier

The internal voltage sources of the amplifier as well as the current sources will frequently be realised by using frequency-dependent elements. As a consequence, the impedances of these sources will vary with frequency, and must in principle be accounted for in the design of the dynamic performance of the bias loop. In anticipation of the discussion on current and voltage source realisations (sections 6.5.1 and 6.5.2, respectively) this seems to be the proper place to point out that a significant influence of these frequency-dependent impedances on bias-loop dynamic performance can easily be avoided.

When the amplifier-stage configurations in the active part are chosen according to the propositions of the preceding chapters, the transfer of the active part is determined by the product of single transfer parameters of the stages. In a cascade of cascode and long-tailed-pair stages, for example, only current-gain factors determine transfer of the active part.

When the bias-voltage sources are realised as bypassed resistive voltage dividers and the current sources as resistors possibly in series with inductances, pairs of poles and zeros are introduced in the transfer function. Even in the case of such simple bias circuitry, however, these poles and zeros are usually so close together that their influence is not noticed in the loop transfer function.

A clarifying example is given in figure 6.6, where part of a possible active part is shown. Bias circuitry is simple. The input stage is a 'folded' cascode loaded with R_1 in parallel with the input impedance of the CB stage. At low frequencies this input impedance is increased by a contribution R_B/h_{fe} (where $R_B = R_2R_3/(R_2 + R_3)$)

because the bypass capacitor C_{bl} is ineffective. A second reason for an increase in the input impedance at low frequencies is that C_{b2} is ineffective as well, so that the CB stage may be loaded with a rather high impedance. Consequently, its internal loop gain may be somewhat reduced.

In spite of these effects, which tend to cause a low-frequency reduction in loop gain, the resistor R_1 can usually easily be made large enough so as to introduce no significant current division, provided that $R_2//R_3$ is not too large. In other words, also at low frequencies the loop gain remains determined by the current-gain factors of the individual stages, and low-frequency loop gain is not significantly reduced. Criteria for the required values of the bypass capacitors obviously do not follow from low-frequency response considerations.

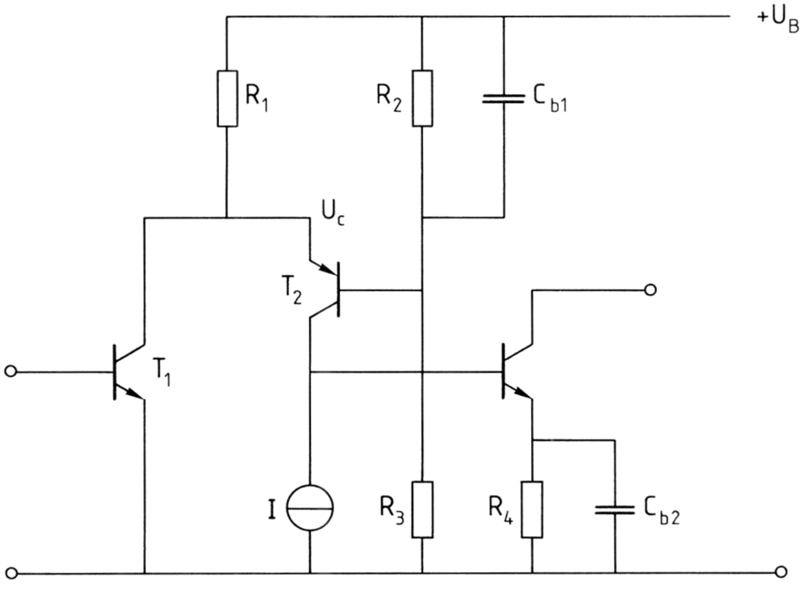


Figure 6.6 Part of an active part of an amplifier with some simple bias circuitry.

We can conclude that, in order to establish the dynamic performance of the biased amplifier, we have to deal only with two or less low-frequency poles in the DC-feedback loop caused by input and/or output coupling capacitors and possibly with one or more poles in the low-pass filter.

The design problem is to realise a stable high-pass filter characteristic of the amplifier, without peaking in the frequency response, or even without ‘undershoot’ in the step response, and with a –3 dB frequency appropriate to the signal information to be processed.

When designing the low-frequency dynamic response of the amplifier it is necessary to bear other quality aspects in mind. That SNR is deteriorated by too small a value of the input coupling capacitor or that the signal-handling capability is reduced by too small a value of the output coupling capacitor is not a hypothetical case. The impedance of the input coupling capacitor acts — when its value is too small — as a series impedance, across which the input-noise current source may cause a significant additional noise voltage at low frequencies [3] (see chapter 3, section 3.3). When the value of the output coupling capacitor is too small, a larger voltage at the output may be necessary for supplying a desired current to the load.

The above phenomena are emphasised here because they are not necessarily accompanied by a low-frequency roll-off of the amplifier transfer function, and are therefore easily overlooked.

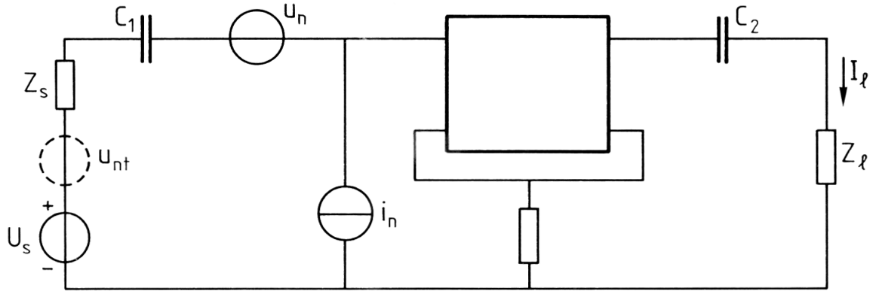


Figure 6.7 Transadmittance amplifier with coupling capacitors at input and output.

Both effects mentioned play a role in a transadmittance amplifier as depicted in figure 6.7, making it an example pre-eminently suited to examining the influence of the bias elements.

1. The spectrum of the total input noise voltage u_{nt} (dashes) is increased by an amount $(1/\omega^2 C_1^2)\ S(i_n)$ as a result of the coupling capacitor C_1. In order to keep this contribution vanishing small, it is reasonable to select a value for C_1 so that $1/\omega C_1 < |Z_s|$ for the lowest frequency to be processed.
2. In order to avoid a substantial signal-voltage drop across C_2, a similar criterion can be used for the value of the output coupling capacitor, viz.: $1/\omega C_2 < |Z_\ell|$ for the lowest frequency.

The requirements following from the desired dynamic performance to be dealt with hereafter may or may not be in agreement with these criteria. If not, the noise and signal-handling considerations may prevail over the low-frequency-response requirements.

As an example of dynamic-response calculation we will discuss the performance of the transadmittance amplifier of figure 6.7 with a bias-feedback loop as given in the previous section (figure 6.5). The following assumptions are made. Source and load impedances are real at low frequencies. The low-pass filter has a first-order transfer function. The parameters A, C and D are zero, while parameter B is real. According to chapter 1, we find for the transfer function $Y'_t = I_\ell/U_s$:

$$Y'_t = \frac{-1}{B + gZ'_sZ'_\ell}$$

where $$Z'_\ell = \frac{1 + p\tau_\ell}{pC_2}, Z'_s = \frac{1 + p\tau_s}{pC_1}, g = \frac{g_o}{1 + p\tau_f}, \tau_\ell = R_\ell C_2, \tau_s = R_sC_1$$

Y'_t can be written alternatively as:

$$Y'_t = \frac{Y_t}{1 + Y_tk}$$

where $Y_t = -1/B$ and $k = -gZ_sZ'_\ell$. The poles of k are *zeros* of Y'_t The roots of the equation $1 + Y_tk = 0$ are the *poles* of Y'_t. In general, a three poles, three zeros response is obtained, but it is possible in this situation to cancel one pole and one zero by taking for example $\tau_\ell = \tau_f$. The loop gain Y_tk can then be written as:

$$Y_tk = Y_t\left(g_o\frac{1 + p\tau_s}{pC_1}\frac{1}{pC_2}\right)$$

In that case, the transfer function Y'_t has two zeros in the origin and two poles following from:

$$p^2 = p\omega_n^2C_1R_2 + \omega_n^2 = 0$$

with $\omega_n^2 = g_oYt/C_1C_2$.

If we require, for example, two coincident real poles for a critically damped high-pass response, it is necessary that

$$\frac{2}{C_1R_2} = \omega_n$$

The step response has a form in this case as shown in figure 6.8.

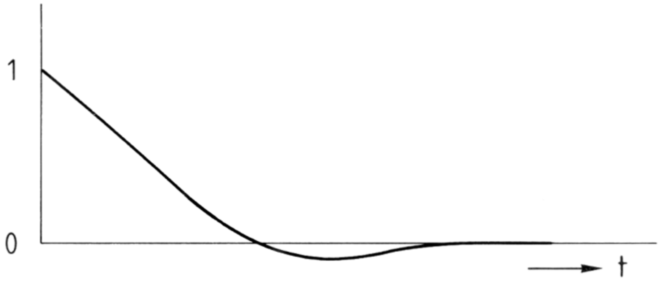


Figure 6.8 Step response of a critically damped second-order high-pass transfer.

The proposed biasing method obviously admits the realisation of a well behaved low-frequency response for this transadmittance amplifier. With other amplifier types similar results can be rather easily obtained. In current and voltage amplifiers, the characteristic equation may even be of the first order, when a pole-zero cancellation similar to the above is accomplished.

Problems arise when source and load impedances have a zero DC value. Pole-zero cancellation is not feasible then, and one may have to resort to biasing schemes using two or more separate bias loops. We will not deal with these in this work.

6.5 Noise and other Contaminating Signals Introduced by the Bias Circuit

6.5.1 Introduction

Practical current and voltage sources produce noise. Moreover, contaminating signals — superimposed on the supply voltage — may penetrate via the bias circuitry into the signal path. For the influence of the bias circuit on signal performance to be negligible, these undesired contributions are required to be small compared to the noise levels inherent in the signal path. We will investigate the performance of some simple current and voltage sources and their applicability as bias components in wide-band amplifiers.

6.5.2 Current sources

In order to compare the performance of various practical approximations of DC current sources we will assume that the current source has to be inserted between the supply voltage U_B and an internal point of the amplifier to which the current must be supplied and which has a fixed voltage U_C.

There are four important aspects regarding the design of DC current sources.

1. The driving-point impedance of the signal path should not be disturbed significantly
2. The current-source impedance must be so large that the influence of signal path noise sources is not significantly increased.
3. The penetration of contaminating signals from the supply voltage into the signal path has to be sufficiently small.
4. The noise production of the current source itself should be well below the noise level inherent in the signal path.

(i) Resistor current source

The most simple approximation of a DC current source is a resistor. An example of its application has already been given in figure 6.6 (R_1). We refer to this figure for an illustrative examination of the various design aspects mentioned above.

The driving-point impedance Z_D in the signal path is determined by the impedance seen in the emitter of T_2. At low frequencies this impedance is approximately given by $Z_D = r_\mathrm{e} + (r_b + R_B)/\alpha_\mathrm{eo}$, where $R_B = R_2 R_3/(R_2 + R_3)$ provided that the load of T_2 is small with respect to r_{ce}. There will usually be no problem in giving R_1 a value large with respect to Z_D which will be at high frequencies even smaller due to C_{bl}.

The driving-point impedance at the collector is much higher. Most likely, a resistor will significantly influence the impedance level and will therefore not suffice.

The resistor R_1 increases the influence of the equivalent input noise-voltage source of the CB stage because the parameters E and p of the input stage (including this output shunt resistor R_1) are reduced. For this reason it will frequently be necessary to eliminate the noise contribution of R_2 and R_3 to this voltage source by using a sufficiently large bypass capacitor $C_{b'}$.

When the supply voltage U_D is assumed to be contaminated by a superimposed noise voltage u_b, while R_1 is large compared to the driving-point impedance concerned, a noise current will be injected in the signal path given by $i_b = u_b/R_1$. This current should be small with respect to the equivalent noise current source at the output of T_1. When T_1 is the first stage, this current source has a value of *at least* $i_c + g_e u_{tb}$, where i_c is the collector shot noise and u_{rb} is the thermal base-resistance noise. The spectrum of this source can be written as

$$S_{io} = 2qI_C\left(1 + \frac{2r_b}{r_e}\right) = 4kT\frac{1}{2r_e}\left(\frac{2r_b}{r_e}\right) \tag{6.3}$$

A worst-case requirement follows for the maximum allowable value of u_b.

The noise current generated by the resistor itself has a spectrum:

$$S(i_R) = 4kT/R_1 + \frac{KI_{DC}^2}{f^\alpha} \tag{6.4}$$

where $\alpha \approx 1$, I_{DC} is the DC current through the resistor and K is a constant, depending on the material, fabrication method and resistance value [4]. The second term in (6.4) obviously represents the excess or $1/f$ noise. Wire-wound and metal-film resistors produce relatively small amounts of excess noise and should therefore be used in critical situations.

For keeping the noise contribution of the resistor R_1 significantly below the noise level of the signal path, it is always sufficient to require that $S(i_R)$ given by (6.4) be small compared to the spectrum given by (6.3).

We have illustrated the design aspects of resistor current sources on the basis of the circuit of figure 6.6, assuming that T_1 was the input stage. In later stages of the amplifier the latter three design aspects may become less important and only the first aspect will usually be decisive for the applicability of a resistor as a current source.

(ii) Resistor-inductance current source

Instead of a resistor one might consider an inductance as a current-source approximation. The use of an inductance alone as a current source, however, would make it impossible to follow the proposed biasing method of sections 6.3 and 6.4 and will not be investigated here. It should be noted that especially for high-frequency amplifiers its use is not entirely unpractical.

We will again assume now that a voltage drop $U_B - U_C$ must be accomplished in an approximated current source formed by a series connection of a resistor and an inductance. We will briefly deal again with the various design aspects.

Compared with the resistor current source and in reference again to figure 6.6, the impedance can be enlarged without changing the voltage drop, which clearly has the advantages that at frequencies above $\omega = R/L$

a) the impedance level of the signal path is less disturbed;

b) the parameters E and a of the input stage (including the ‘current-source load’) are reduced less, so that the influence of the second-stage noise voltage source is smaller;

c) the noise current penetrating into the signal path caused by the noise voltage u_b is smaller.

A further advantage is that the noise current of this current source is smaller. The thermal noise is given now by:

$$S(i) = 4kT \frac{R}{R^2 + \omega^2 L^2}$$

The excess noise will not change significantly. A disadvantage may be that practical low-cost inductances are sensitive to external fields, which may cause an R.F. interference-current in the signal path.

(iii) Active current sources.

For a given DC current value and voltage drop, the impedance of an active current source can be very much higher than that of a resistor, at least at low frequencies. We will only deal here with the extremely simple current sources using one active device. Such a bipolar-transistor current source is depicted in figure 6.9.

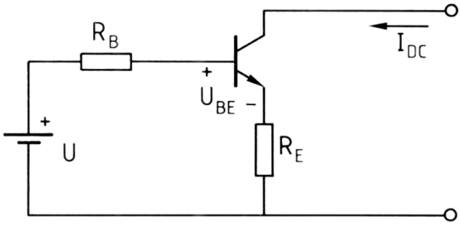


Figure 6.9 Simple bipolar-transistor current source.

The DC current generated by this current source is approximately given by

$$I_{DC} = \frac{U - U_{BE}}{R_E + R_B/h_{FE}} \tag{6.5}$$

provided that the collector-base junction is reverse biased.

Many textbooks (e.g. [5]) deal with the output impedance of such current sources. We refer to these for exact expressions. Here we will ascertain that the maximum output impedance is obtained when $R_B \to 0$ and $R_\mathrm{E} \to \infty$. It is approximately composed in that case by a shunt connection of a resistance $\alpha_\mathrm{eo} r_\mathrm{ce}$ and the capacitance $C_{b'\mathrm{c}}$. In practice, we can bypass the base resistors and have R_E as large as is consistent with the requirements following from (6.5) and the reverse-biased collector-base junction.

Only with very high driving-point impedances in the signal path will such a current source significantly influence the impedance level. Usually, only the output capacitance will then be relevant. In cases where a reduction in this output capacitance is required, a second active device may be used to realise a cascode configuration according to figure 6.10. The output impedance has now increased, because a current flowing into $C_{b'c}$ is largely returned through the transistors.

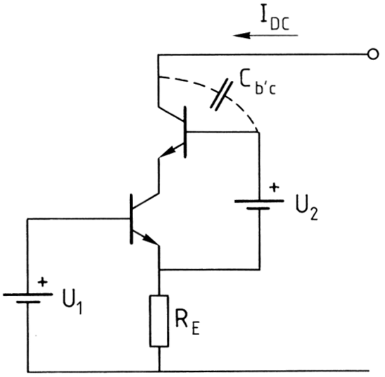


Figure 6.10 Cascode current source.

As far as the first two design aspects of current sources as formulated earlier are concerned, active current sources perform much better than the passive current-source approximations.

Now, we will discuss the latter two design aspects. First, we will turn to the noise production of the simple current source of figure 6.9.

The voltage source U is assumed not to produce any noise. The spectrum of the output-noise current for not too high frequencies can be written as:

$$S(i_{DC}) \approx 4kT\left[\frac{(r_b + \frac{1}{2}r_e + R_B + R_E) + \frac{1}{2h_{FE}r_e}\left(1 + \frac{f_\ell}{f}\right)(r_b + R_B + R_E)^2}{\left(\frac{r_b + R_B}{\alpha_e} + r_e + R_E\right)^2}\right]$$

where the excess noise of the resistors is disregarded.

The spectrum has a minimum for $R_B \rightarrow 0$ and $R_E \rightarrow \infty$ given by

$$S(i_{DC})_{min} = 2qI_B(1 + f_\ell/f)$$

f_ℓ was defined in section 3.7.2.

The condition $R_B \rightarrow 0$ can be met by bypassing the base with a sufficiently large capacitor. In practice, the noise contribution of R_E will generally dominate, because large resistor values ($R_E \gg\cdot 2h_{FE}r_e$) require large supply voltages. For $h_{FE} = 100$, for example, the contributions of the thermal noise of R_E and of the base shot noise are equal when the voltage drop across R_E amounts to 5 volts.

With moderate voltage drops across R_E, the output current is determined by the spectrum of R_E predominantly. Because some voltage drop is required across the transistor as well, the noise production of a transistor current source always exceeds that of a resistor. To limit the noise production at high frequencies, a coil can be inserted in series with the emitter resistor.

FET current sources in principle offer the possibility of obtaining better noise performance because of the virtual absence of an input noise current.

In simple current sources, the reference voltage U (figure 6.8) may be realised with a resistive divider that is bypassed by a capacitor in order to enlarge the output impedance and to reduce the noise. This bypass capacitor may be either connected to ground (the common terminal of the amplifier, signal source and load), or to the supply voltage. The supply voltage is assumed to be contaminated by an AC voltage ub with respect to ground. For the determination of the current injected in the signal path, due to this voltage, we can use the signal circuits of figure 6.11a and 6.11b, showing respectively the situations where the base is bypassed first to the supply voltage and then to the ground.

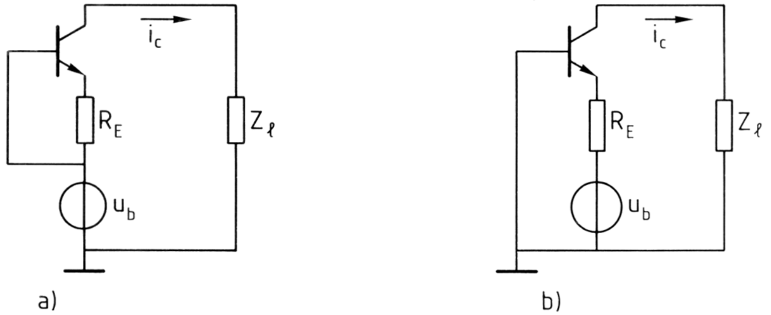


Figure 6.11 Base of a current source bypassed to supply voltage (a) and to ground (b).

In figure 6.11a, the current source behaves as a one-port with respect to the voltage u_b. The injected current i_c is equal to u_b/Z_o, where Z_o is the output impedance of the circuit. For a sufficiently large value of R_E ($R_E \gg h_{FE}r_e$ or $U_{RE} \gg 2.5$ V at $h_{FE} = 100$), the current-source impedance is formed by about $\alpha_{eo}r_{ce}$, shunted with $C_{b'c}$. The penetration of low-frequency supply-voltage variations can be extremely small. At higher frequencies the transfer from u_b to i_c is determined by $C_{b'c}$.

When the bypass capacitor is connected to ground, the current flow due to u_b is determined by R_E. High-frequency rejection can be increased then by inserting a coil in series with R_E.

The value of the bypass capacitor in the base lead usually has to be selected on the grounds of noise performance and supply-voltage rejection.

Better temperature stability of the current source can be realised with the aid of compensation techniques. We refer to the comprehensive literature on this subject [6].

6.5.3 Admissible noise contributions of the bias current source at the input

The most critical point where a bias current source may be needed is at the input of an amplifier. A large resistor which is connected to an output-voltage-controlled source may be used in many situations, instead of the controlled current source of figure 6.5. In order to have a noise contribution which is small with respect to the equivalent input noise source, the resistor must have a certain minimum value which will be established below.

If a bipolar transistor is used in the amplifier input stage, the minimum input noise current amounts to $2qI_B \approx 4kT/2h_{FE}r_e$. The current I_B, flowing through the bias resistance R_B, causes a voltage drop $U_{RB} = I_BR_B$. The thermal-noise spectrum of this resistance is given by $S(i_{RB}) = 4kT/R_B$. In order that this spectrum be small with respect to $2qI_B$, it is necessary that:

$$R_B >> 2h_{FE}r_e$$

or $$U_{RB} \gg 2h_{FE}I_B\frac{kT}{qI_E} \approx \frac{2kT}{q} \approx 50 \text{ mV}$$

at room temperature. A similar condition is found for the gate leakage resistance of an FET. Generally, this voltage drop will be small enough to guarantee sufficiently accurate biasing.

For obtaining a large input impedance in series-feedback situations, higher values of R_B may be necessary. As an attractive alternative that maintains accurate biasing, a bootstrapping technique may be used, which is schematically shown in figure 6.12 for a transadmittance amplifier. The influence of the bias-feedback loop on signal performance is made very small by the elimination of the negative feedback, by means of the bypass capacitor C_B. There is virtually no signal-current flow in R_B. Some positive feedback is introduced at high frequencies, due to the current supplied via R in the non-inverting input. The output impedance of the amplifier may become negative as a consequence.

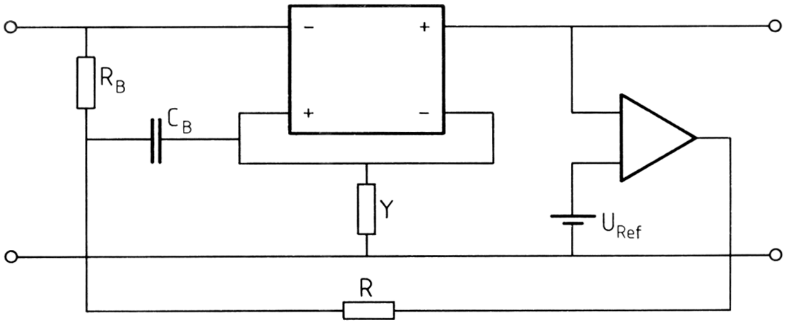


Figure 6.12 Bootstrapping for preventing input impedance reduction by the bias resistance in the case of input series feedback.

In the case of an FET input stage, much larger bias resistances are needed because of the extremely low value of the gate leakage current. Values up to 1 GΩ [2] or forward-biased diodes [7] may be used.

6.5.4 Voltage sources

The reference voltages, used to establish the value of U_{CB} or U_{DG}, of the various active devices may be realised by bypassed resistive dividers, or with the aid of low-differential-impedance devices such as diodes, zener and avalanche diodes, U_{BE} multipliers and voltage followers, whether or not in combination with bypass capacitors used to reduce the impedances and noise production at higher frequencies. We will not deal with these realisations here, but refer instead to the literature [6]. Design aspects similar to those for current sources are involved.

Weighing the importance of these aspects for various implementations and applications is left to the reader.

6.6 Discussion

In this chapter, we have presented a formal approach to the design of bias circuitry for negative-feedback amplifiers that do not have to process very low-frequency signals. Theoretically, the bias components needed are ideal (controlled) current sources, ideal voltage sources, low-pass filters and high-input-impedance differential-input amplifier. When these components are used, the active devices of the signal path can be biased individually.

Alternatively and preferably, all amplifier stages take part in the bias circuit of a certain stage. For each stage in the amplifier circuit, a bias circuit with a well-defined structure must then be demonstrable.

The influence of the bias circuit on the signal performance of the amplifier can be modelled with the aid of Black's elementary feedback model. Bias components within the active part of the amplifier will generally have no significant influence on the dynamic response because poles and zeros are nearly coincident in pairs when the active part has a structure as proposed in the previous chapters.

The dynamic low-frequency response can easily be designed, unless source and load impedances have a zero DC resistance. It may be necessary in that case to use two separate bias loops.

The values of the coupling capacitors at the input and output, which are frequently needed in order to prevent DC current from flowing through the source or load, frequently have to be selected on the grounds of noise performance and signal-handling capability rather than on the desired low-frequency response.

The current sources and voltage sources needed for biasing purposes have to be designed so that the intended amplifier performance is not significantly disturbed. Practical design aspects for current sources have been discussed in some detail.

References

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7 Outline of the Design Method

7.1 Introduction

In this final chapter we will give an outline of the design method as it was unfolded and justified more or less explicitly in the previous chapters.

It may be useful to recapitulate the restrictions we have made in this work.

First, we have dealt exclusively with negative-feedback amplifiers having one terminal in common with the source and load, because a disturbance, by contaminating signals and parasitic impedances, is much more likely to occur in the case of floating sources and loads.

A second restriction was imposed on the type of information to be processed. Very low frequencies are assumed to be absent in the relevant information, so that slow variations of the bias quantities due to temperature or supply-voltage variations can be distinguished from the signal information. This means that so-called DC amplifiers, where some additional design aspects have to be taken into account — were excluded from the discussion. The quality aspects that were discussed in this work refer also to DC amplifiers, of course.

Selective amplifiers were excluded from consideration as well.

Along with the restrictions we made, some starting points were formulated in chapter 1 that were justified in later chapters. The most important factors put forward there are the necessity of feedback for obtaining an accurate and linear transfer function, and — in order to avoid SNR deterioration or output-power loss — the omission of series or shunt impedances at the input and output.

In this chapter we will combine the conclusions of earlier chapters as far as necessary in order to elucidate and explain the design procedure.

7.2 The Selection of the Proper Basic Amplifier Configuration

7.2.1 Impedance adaptation to the source and load

The basic amplifier configuration — needed to solve a particular amplifier design problem — has to be selected on the grounds of the properties of the signal source and load.

The signal source produces information in response to a physical quantity in the form of a current, a voltage, a charge, etc. That electrical quantity that has the best linear and accurate relationship to the primary signal is decisive for the selection of the nature of the amplifier input impedance.

When, for example, a *current* -— accurately and linearly related to the primary physical signal — is produced in the source, a *low* input impedance is required, in order to prevent the possibly inaccurate or non-linear source impedance from playing a role in the information transfer.

A *voltage*-producing signal source requires a *high* input impedance for similar reasons.

When the source impedance is *linear* and *accurately known,* a linear and accurate input impedance *can* be used. Accurate impedance matching is frequently *needed* in characteristic-impedance systems in order to avoid standing waves. Similar considerations are valid with respect to the desired nature of the output impedance.

The important consideration that the impedance adaptation should not be realised by the use of series or shunt impedances, because — as shown in chapters 3 and 4 — these impedances lead to a deterioration of the SNR and power efficiency in wide-band amplifiers, leads to a preference for setting the desired impedance values with the aid of negative feedback.

Because negative feedback is indispensable for obtaining a transfer independent of parameter variations in the active devices, both the transfer stabilisation and the impedance adaptation must be realised by the proper type(s) of negative feedback.

7.2.2 Classes of negative-feedback amplifiers

In order to find the proper amplifier type for a given application we have to know all possibilities for realising amplifiers with fundamentally different transfer properties. For this purpose it is necessary to make an inventory of classes of amplifiers.

(i) Non-energic-feedback amplifiers

Chapter 1 presents a classification of *non-energic* negative-feedback amplifiers with up to four overall feedback loops. Non-energicness guarantees that there is no deterioration of the SNR and power efficiency due to the feedback networks. The classification encompasses the complete set of twoport amplifier types with fundamentally different transfer properties.

(ii) Transformer-feedback amplifiers

Thereupon, similar amplifier types are presented with transformer feedback. The designer has thus at his disposal a set of amplifier types that is most suitable for wide-band applications, including characteristic impedance matching with small SNR degradation and output-power loss. These amplifiers are not suited to processing (very) low frequencies.

(iii) Impedance-feedback amplifiers

The subsequent abandonment of transformers in the feedback networks reduces the number of realisable amplifier types with direct sensing of the output quantity and direct comparison of the input and feedback quantities. Moreover, it necessitates the use of active amplifier parts with floating ports, when series feedback is required. This may lead to some inaccuracy and non-linearity even with infinite loop gain as discussed in chapter 4.

In some critical situations, where the source or load have very high impedances, special measures should be taken in order to avoid these errors.

(iv) Active-feedback amplifiers

Compensation for the loss of amplifier types is found in a class of amplifiers with active feedback networks. Two types of active feedback can be distinguished. The first uses indirect sensing of the output quantity and/or indirect comparison of input and feedback quantities. It is designated as an indirect-feedback technique. The indirect-feedback amplifiers are inferior to their direct-feedback counterparts. They are useful configurations, however, for realising some special transfer functions and, because they rely on well-matched devices, they are especially suited to application in integrated circuits.

The second is called an active-feedback technique. Though the feedback network comprises one or more active devices the sensing and comparison at the output and input are direct. This technique also offers the possibility of realising some special transfer functions, similar to those obtained with indirect feedback.

With the aid of the above classifications, an abundance of amplifier types comes to be at the disposal of the amplifier designer. For each combination of source and load, the most appropriate basic configuration can be selected.

(v) Influence of feedback network on noise performance

In all but the non-energic feedback situations, the influence of the feedback network on noise performance cannot be completely disregarded. Chapter 3 provides simple means for taking this influence into account.

7.2.3 Modeling of negative-feedback amplifiers

The design of negative-feedback amplifiers requires an appropriate feedback model. The rather unusual asymptotic-gain model describes the transfer function of an amplifier by its ideal value (following from a nullor approximation of the active part of the amplifier, and designated as $A_{t\infty}$) times a correction factor, which is determined by the loop gain $A\beta$. The design of the basic configuration is concerned with the asymptotic gain $A_{t\infty}$ and in the case of accurate input or output impedances with their asymptotic values Z_∞.

The design of the loop transfer function A is treated in chapters 2-5. This part of the design is concerned with the optimisation of the noise, distortion and accuracy performance and includes high-frequency performance.

7.3 Characterisation of Stages for use in the Loop Transfer Function

The designer, in order to be able to make a justifiable selection of amplifier stages for the active part, should have a complete survey of possible configurations. The configurations of the stages that can be considered for practical use in the active part of the amplifier are given in chapter 2.

A classification of stages can be given similar to that of chapter 1. The CE stage (bipolar transistor) and the CS stage (field-effect transistor) are regarded as stages without feedback. When they are identified with a nullor, the other types of configurations with a single active device are feedback versions of the CE and CS stages.

The negative-feedback stages are considered to be practical only when they can be realised without transformers. Positive-feedback versions cannot be realised without transformers or active feedback networks and are excluded. Current mirrors and current-gain cells are regarded as stages with indirect local feedback.

Balanced versions can be made of all configurations. These have, apart from a better linearity and a somewhat worse noise performance, very similar transfer properties.

The stages are for investigating their transfer properties preferably described by transmission parameters or by their reciprocal values, the transfer parameters. A striking feature of this representation is that a certain type of feedback essentially affects only one transfer parameter. In a stage with series feedback at its output and shunt feedback at its input (CB stage), for example, the current-gain factor is the transfer parameter stabilised by the feedback action. The other transfer parameters very nearly equal those of the CE stage.

Qualitative design considerations can fruitfully be based on the transmission parameter characterisation of the stages. Quantitative design aspects can better be approached from a more physical representation of the stages. The hybrid-π equivalent circuit is used to represent bipolar-transistor performance. A similar equivalent circuit is used for FET's.

The classification and characterisation of single-device configurations given in chapter 2 are used to formulate design criteria in chapters 3, 4 and 5.

7.4 Selection and Design of the Input Stage

The first design step with respect to the loop transfer function concerns the selection of a proper input stage, given the signal-source impedance and the type of information-representing quantity *produced* by the source.

The design considerations are aimed at the minimisation of the contribution of the amplifier noise to the noise produced by the signal source. The input stage produces a very important noise contribution. Therefore, the noise performance of the various possible input configurations has to be compared first.

7.4.1 Noise performance of the single-device configurations

In order to judge the noise performance of the various single-device configurations, their equivalent input noise sources can be qualitatively compared with those of the CE stage. The impedances in the feedback networks may (slightly) deteriorate noise performance, not only by producing noise themselves, but also by enlarging the influence of the active-device noise sources. The noise sources of the CC and CB stages-being essentially non-energic local-feedback stages — are very nearly equal to those of the CE stage.

7.4.2 Influence of the second-stage noise sources

Even when their feedback is non-energic, local-feedback configurations are not recommended for use in the input stage, because at least one of their transfer parameters is reduced. As a consequence, the influence of the second-stage noise is enlarged. The stages without negative feedback, i.e. the CE and the CS stages, are preferable as input configurations for this reason.

7.4.3 Optimisation of noise performance for a given signal source

The design procedure presented in chapter 3 optimises the noise performance of an amplifier for a given signal source, assuming that the noise contribution of the amplifier is determined by the noise sources of the active device in the input stage (CE or CS stage).

The power-density spectra of the equivalent noise sources in series or in parallel with the signal sources are calculated for bipolar transistors as well as for FET's in the case of real, capacitive and inductive source impedances. The total noise can then be calculated by integrating the spectrum over the relevant frequency band.

This total noise can be minimised by selecting a device with favourable parameters and by choosing the proper bias current. This noise minimisation implies the optimisation of the signal-to-noise ratio or noise figure.

Instead of minimising the total noise, the weighted spectrum can be integrated and minimised, or the spectrum itself can be minimised at a certain frequency. Procedures for these alternative optimisations are all very similar to the one dealt with in chapter 3.

7.5 Selection of the Other Stages in the Active Part

7.5.1 Distortion and accuracy considerations

The next step in designing the loop transfer function $A\beta$ may concern the selection of the output-stage configuration. A prerequisite for this stage is that its bias conditions should be such that the desired output-signal-handling capability can be obtained.

The asymptotic-gain model shows very directly the way to realise high accuracy and low distortion. As a reminder, note that this model refers to the transfer function that expresses the *load* quantity in the quantity *produced* by the source. Input and output impedances of the amplifier need not be known explicitly in the case of single-loop amplifiers.

For high accuracy and low distortion, the loop gain $A\beta$ should be as large, as linear and as accurate as possible everywhere in the frequency region of interest. Accuracy and linearity, however, can be exchanged for loop-gain magnitude, and the differential-error-to-gain ratio provides means for estimating the merits of such an exchange.

Non-linearity of the amplifier is very likely determined by the non-linearity of the output stage. For reducing the influence of this non-linearity, loop gain cannot be made arbitrarily large, because of the high-frequency behaviour of the active devices. The number of stages is restricted.

(i) Distortion performance of the single-device configurations

The selection of the most appropriate output-stage configuration is based on a comparison of configurations with the CE stage. It is shown in chapter 4 that the insertion of series and shunt impedances at the input or at the output of the output stage leads to an increased value of the D-G ratio. Local-feedback stages with impedances in their feedback networks have increased D-G ratios as well. The application of non-energic local feedback does not change the D-G ratio essentially.

(ii) Influence of the driving-stage distortion

The presence of local feedback in the output stage leads to a reduction in the value of at least one of its transfer parameters, and requires therefore an increase in the output level of the driving stage in order to maintain the same output level.

This output level is set in an overall negative-feedback amplifier by the amplifier-input level and by the asymptotic gain. As a result of the enhanced driving level, distortion is very likely increased, because the distortion in the driving stage can no longer be disregarded.

More or less accidental compensation effects may lead to a reduction in a certain harmonic component. Only those compensations that make use of well-known opposite variations of equal parameters are considered to be reliable. Some examples were given in chapter 4.

The CE stage is, on the grounds of the above considerations, regarded as the best configuration for use in the output stage.

7.5.2 High-frequency performance

Noise, accuracy and linearity optimisation is achieved by using as many stages as possible in the active part, each having transfer parameters as large as possible. Consequently preferably only non-feedback stages are used in the active part of the amplifier.

The necessity for a well-controlled high-frequency roll-off of the amplitude characteristic of the amplifier leads to additional requirements that may impede the realisation of this preferred structure of the active part. The extent to which this impediment exists depends on the desired asymptotic-gain value, on the way in which it is realised and on the natures of the source and load. When so-called phantom zeros can be used (for shaping the root locus in such a way that the poles of the amplifier transfer function can be directed into their desired positions) the structure of the active part may be rather arbitrary.

(i) Description of the design problem

The design problem for optimum accuracy and linearity combined with a prescribed type of high-frequency roll-off can be summarised as follows. Such optimisation requires a maximisation of the area in the Bode diagram bounded by the 0-dB line and the asymptotes of the loop transfer function under the restriction that the amplifier must have a well-defined filter characteristic (in addition to a possibly intended frequency dependency).

The Bode diagram of the loop transfer function may have a horizontal asymptote for low frequencies. There is no reason to enlarge the loop gain below the lowest frequency to be processed. So the first pole of the loop transfer function may correspond to the lowest frequency in the signal. One might consider in some special cases the application of local positive feedback for enlarging the low-frequency loop gain. Base-current compensation [1] is an example of such a technique. Generally, however, positive feedback must be considered as having little practical significance.

The above-mentioned area in the Bode diagram can be larger, as the active devices have larger transit frequencies.

As a vehicle for high-frequency-performance considerations, we have chosen the MFM all-pole filter approximation, because its parameters are simply related to bandwidth.

(ii) Bandwidth capabilities of single-device configurations

As a figure of merit for bandwidth capabilities of a stage, the contribution of this stage to the product of low-frequency loop gain and magnitudes of the poles — the so-called L-P product — was introduced. The more familiar classical G-B product may be a misleading quantity for this purpose.

It is shown in chapter 5 that all single-device configurations can roughly provide equal contributions to the L-P product and that series and shunt impedances in the signal path may deteriorate these contributions.

(iii) Designability of the loop transfer function

Good designability either requires the absence of interaction between stage transfer functions or the presence of a well-known interaction.

Interaction can be reduced or controlled in various ways. The insertion of series and shunt impedances in the signal path is one of them. The use of local feedback, however, provides more a preferable means of achieving less deterioration of noise and distortion performance.

Of the various possible combinations of single-device configurations, the cascode (inverting) and the long-tailed pair (non-inverting) are pre-eminently suited to reducing the interaction. The initial active-part design can be done with these unilateral stages. Though these stages are cascades of two configurations, viz. of a CE and a CB stage and a CC and a CB stage, respectively, they can be regarded as virtually unilateral single-device configurations. They frequently contribute — under most drive and load conditions — one significant pole to the loop transfer function. Only this pole will arrive in the bandwidth-determining dominant group of the amplifier transfer function. The effective contribution to the L-P product of these stages amounts to WT

In order to produce the desired well-controlled interaction, several techniques — known as compensation techniques — can be used. Some of these are described in chapter 5. Because of reasons given before, the techniques using local feedback are preferred to those that use shunt or series impedances. The required amount of local feedback can usually be obtained by replacing a unilateral stage with a proper single-device stage.

(iv) Design procedures

Chapter 5 gives design procedures for two-poles and three-poles transfer functions and various numbers of available phantom zeros. Moreover, the influence of an extra pole (or of excess phase caused by several distant poles) is investigated. The sensitivity to excess phase appears to get smaller as more phantom zeros are used. In practice, the use of phantom zeros will frequently introduce extra more or less remote poles.

7.5.3 Some comments on intermodulation distortion

At this point it would appear to be quite useful to discuss one specific quality aspect — namely, intermodulation distortion — that was not mentioned before explicitly. The considerations also refer to the phenomena of slewing and transient-intermodulation distortion.

Every amplifier is designed for processing the information in a restricted frequency range. However, interfering signals having large amplitudes at high frequencies but lying outside the relevant band may be present in the input signal. In order to avoid intermodulation distortion or slewing phenomena (that may bring undesired frequency components in the relevant band) the amplifier must be capable of linearly processing these interfering signals as well, unless they can be filtered adequately before they reach the amplifier.

Such filtering is always useful to relieve the linearity requirements. In some cases, such as radio-frequency interferences caused by careless wiring, for example, screening may be sufficient. When the interfering signals are inherent in the signal source, filtering networks are required, which may adversely influence noise performance.

If no adequate input filtering can be achieved, all amplifier stages must have a sufficient signal-handling capability, whether they are part of an overall negative-feedback amplifier or not.

When amplifiers are designed such that they must be stable for all kinds of conditions of the source, load and feedback network such as internally compensated operational amplifiers, it is impossible for the performance in this respect to be optimum, because the L-P product is necessarily reduced to a large extent.

When amplifiers are designed so that each stage contributes its maximum amount to the L-P product, (cascaded) overall negative-feedback amplifiers will give better results than cascaded local-feedback stages. Additional measures to relieve signal-handling requirements are found in a reduction of the high-frequency asymptotic gain, so that the load-power level is reduced. Consequently, the drive requirements are reduced as well.

It should be emphasised that filtering *within* the active part of the amplifier has an adverse effect on all types of distortion mentioned.

7.6 Bias-Circuit Design

The final part of the design is concerned with the bias circuit. Because the design of the signal path is completed first, the bias circuit is not allowed to interfere with signal performance. Various kinds of interference are possible. Some of them were not discussed as a consequence of the restrictions we made on the types of amplifiers to be dealt with. We excluded very low-frequency bias-current and voltage variations from consideration because they can be separated from the signal by proper filtering.

7.6.1 Design aspects

The various design aspects involved in the design of bias circuitry are listed below.

- Setting of the bias currents and voltages of all active devices.
- Dynamic performance; low-frequency response.
- Influence of the bias-circuit components on the impedance levels of the signal path.
- Generation of noise in the bias components.
- Penetration of noise and other contaminating signals from the supply voltage via the bias circuit into the signal path.
- Enhancement of the influence of the noise sources of the active devices and reduction of the signal-handling capability caused by non-zero impedances of bypass and coupling capacitors and by finite impedances of RF chokes.

7.6.2 Design approach

Chapter 6 presents a formal approach to basic bias-circuit design with respect to static and dynamic performance. Each stage has to be provided with two reference sources setting the collector-to-base voltage and the collector current (in the case of a bipolar transistor and the corresponding quantities in the case of an FET). A controlled source is used at the input of the stage in order to fix these quantities via a feedback loop, which generally has to contain a low-pass filter. This low-pass filter is needed to maintain the intended impedance levels in the signal path.

The reference current and voltage sources can be derived from the supply voltage. The bias-feedback loop of a stage usually comprises all other stages of the amplifier.

7.6.3 Dynamic performance

The dynamic performance of an amplifier with an overall bias-feedback loop must be carefully designed by giving the proper values to the poles and zeros. These values are determined by source and load impedances and by the coupling capacitors needed to prevent bias currents from flowing through the source and load. Moreover, one pole is present in the transfer function of the low-pass filter.

In some cases, the bias circuit must be split into two parts because the source and/or load impedances may have characteristics such that no satisfactory dynamic performance can be realised.

7.6.4 Influence of bias circuitry on the signal-path performance

The non-zero impedances of coupling and bypass capacitors and the finite impedances of RF chokes may adversely influence the noise performance and signal-handling capability in the low-frequency region. The requirements regarding these aspects may prevail over those of dynamic performance.

The noise generated in the bias components must be small compared with the noise sources of the signal path. The impedances introduced by the bias circuit must be either large (when they are shunting the signal path) or small (when they are in series) with respect to the impedance levels on the spot.

The noise and other contaminating signals penetrating from the supply voltage into the signal path via the bias circuit must be kept small by minimising their transfer.

The above requirements refer to the practical implementation of approximate current and voltage sources. Some examples and criteria were given in chapter 6.

7.6.5 Concluding remarks

Detailed design of a bias circuit is largely a matter of available technology. Wide-band transformer-feedback amplifiers with discrete components, for example, are very different from bipolar integrated circuits, as far as the implementation of reference sources is concerned, as well as with respect to the types of components in the signal path. Each method of implementation requires its own artistry in addition to a systematic design approach as presented in this work.

7.7 Examples of the Design Procedure

During the course of this work many amplifiers were designed according to the design method presented. A short description together with a reference to the reported results will be given for some amplifiers. The fruitfulness of the design method is demonstrated by the outstanding qualities of the amplifiers.

7.7.1 Preamplifiers for moving-coil phono cartridges

A preamplifier is described for transducers with very low source impedances, typically < 10 ohms, producing a very small voltage (< 100 μV) in response to the primary information-representing quantity. In the case of a moving-coil phono cartridge this quantity is the needle velocity.

In order to realise an acceptable signal-to-noise ratio, the input (bipolar) transistor must be carefully designed. A special transistor with a very low base resistance of about 1.5 ohms has been fabricated [2].

First, a simple amplifier with discrete components was realised with this input transistor [3], and then a completely integrated version was designed with a balanced input stage [4]. In the latter case a base-current compensation could be employed such that a large coupling capacitor at the input could be avoided. In spite of the balanced input stage an adequate signal-to-noise ratio could be realised.

7.7.2 Active-antenna circuits

For realising a wide-band antenna-matching network for the frequency range from 5 kHz to 30 MHz various amplifier types can be considered.

Short antenna rods ($\ell \ll \lambda$) have capacitive source impedances. A frequency-independent response to the electric component of the electro-magnetic field requires a frequency-independent amplification of the open antenna voltage or an integration of its short-circuit current. Because of the high source impedance and the need for protecting the input active device, an amplifier with a low input impedance is preferred, as it minimises the influence parasitic impedances at the input have on the transfer quality.

For optimum noise performance a junction FET at the input should be used.

For optimum distortion performance and signal-handling capability, two feedback loops can be used, resulting in an accurate output impedance of 50 ohms.

Active antennas covering the complete frequency range of 5 kHz - 30 MHz with a length of 50 cm are described in [5]. They exhibit excellent noise and intermodulation performance and are thoroughly protected against static. Output power levels up to +28 dBm can be handled which correspond to electric-field strengths of over 10 V/m.

An integrated amplifier with a single feedback loop for an active-car-radio antenna is described in [6]. It was first necessary to develop a new method for realising a junction FET in a standard bipolar process [7]. The noise and intermodulation performance of the amplifier with a 30 cm whip antenna are comparable to those of a car radio with a 1 m antenna.

7.7.3 Camera-tube preamplifiers

The design of current amplifiers suitable for amplifying the signal of camera tubes or other similar transducers (for example, photodiodes) is described in [8].

The advantage of using a current amplifier rather than the more conventional transimpedance amplifier lies in the possibility to use much larger feedback resistances. Especially at low frequencies the noise performance can be improved while simultaneously obtaining a very large bandwidth (> 10 MHz).

Experimental results are presented for a camera tube with a 15 pF output capacitance. The measured equivalent input noise current of this amplifier is smaller than 1 nA in a 5 MHz bandwidth. For obtaining this outstanding noise performance it is necessary to use two high-ω_T junction FET's in parallel at the input.

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Appendix

Influence of the Induced Gate Noise in Wide-Band Amplifiers

The thermal channel noise of an FET is represented by two correlated current sources, according to figure A.1, where i_1 is the induced gate noise. The spectra of i_1 and i_2 are given by:

$$S(i_1) = 4kT\frac{\omega^2 C_{gs}^2 a}{g_m} \tag{A.1}$$

$$S(i_2) = 4kTbg_{\mathrm{m}} \tag{A.2}$$

and $$S(i_1 i_2) = \gamma S^{\frac{1}{2}}(i_1) S^{\frac{1}{2}}(i_2) \tag{A.3}$$

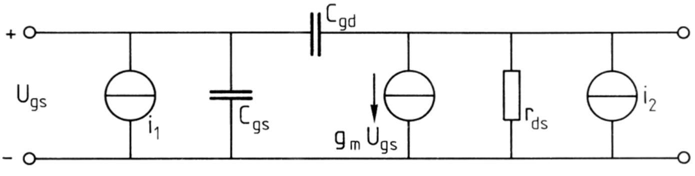

Figure A.1 Thermal noise sources in a junction FFT.

where g_m = transconductance, $\frac{2}{3} C'_{gs}$, in which C'_{gs} is the capacitance of the gate area, and γ is the complex correlation factor defined by (A.3) in which $S(i_1 i_2) = \langle I_1^* \, I_2 \rangle$.
According to Klaassen [1]:

$$a = \frac{1}{3}, \quad \frac{2}{3} < b < 1$$

$$\gamma = \gamma_r \frac{\omega}{\omega_1} + \mathrm{j}\gamma_i$$

with $0.10 < -\gamma_r < 0.15$, and $0.30 < \gamma_i < 0.40$, and $\omega_1 = g_m/bC_{gs}$.
The transformation of i_2 to the input, together with i_1, gives the equivalent noise sources at the input U and I:

$$U = \frac{I_2}{g_m} \quad \text{and} \quad I = I_1 - I_2 \frac{\mathrm{j}\omega(C_{gd} + C_{gs})}{g_m} \tag{A.4}$$

If this FET is connected toa source with admittance $y_s = g_s + jb_s$, the equivalent noise current source I_e at the input (figure A.2) has an energy density spectrum $S(i_e)$:

$$S(i_e) = \langle (I \angle Y\psi_\sigma)(I - Uy_s)^* \rangle$$

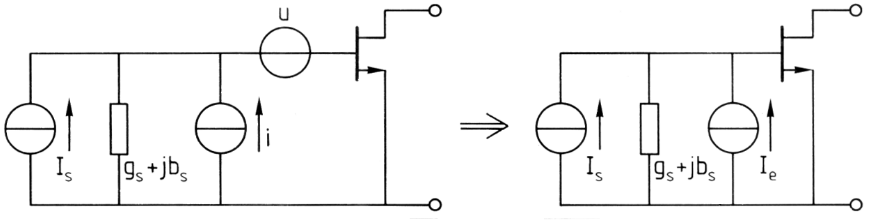

Figure A.2 Transformation of input noise sources into equivalent current source I_e.

We obtain with (A.3) and (A.4):

$$S(i_e) = \frac{S(i_2)}{|A|^2} \left\{ 1 + |A|^2 \frac{S(i_1)}{S(i_2)} - 2\,|A| \frac{S^{\frac{1}{2}}(i_1)}{S^{\frac{1}{2}}(i_2)} \operatorname{Re} \gamma \frac{A^*}{|A|} \right\} \tag{A.5}$$

in which

$$A = \frac{g_m}{y_s + j\omega(C_{gs} + C_{gd})}$$

Expression (A.5) has a form:

$$S(i_e) = \xi\{1 + \lambda^2 - 2\delta\lambda\}$$

with

$$|\,\delta| = |\operatorname{Re} \left\{ \gamma \frac{A^*}{|A|} \right\}| \le 1$$

it can easily be shown that this expression has a lower boundary

$$S(i_e) \ge \frac{1}{|A|^2}\; S(i_2) \left\{ 1 - |A| \frac{S^{\frac{1}{2}}(i_1)}{S^{\frac{1}{2}}(i_2)} \right\}^2$$

and an upper boundary

$$S(i_e) \le \frac{1}{|A|^2}\; S(i_2) \left\{ 1 + |A| \frac{S^{\frac{1}{2}}(i_1)}{S^{\frac{1}{2}}(i_2)} \right\}^2$$

From these boundaries it can be seen that the induced gate noise can be neglected if:

$$|A|\frac{S^{\frac{1}{2}}(i_1)}{S^{\frac{1}{2}}(i_2)} \ll 1 \quad \text{or} \quad |\mathrm{j}\omega C_{gs} + \mathrm{j}\omega C_{gd} + y_s| \gg \omega C_{gs}\sqrt{\frac{a}{b}}$$

In wide-band amplifiers this condition is always met.

Reference

1 F.M. Klaassen, IEEE Trans. on Electron Devices, 14 (1964) 368-373.

List of Symbols

A	gain (unspecified)
A	$(U_i/U_o)_{I_o=0}$, transmission parameter
A_i	current gain
A_t	gain (unspecified) of a feedback amplifier
$A_{t\infty}$	asymptotic gain of a feedback amplifier
A_u	voltage gain
$A\beta$	loop gain
$A_o\beta_o$	Low-frequency loop gain
B	$(U_i/I_o)_{U_o=0}$, transmission parameter
$B(\omega)$	bandwidth
C	$(I_i/U_o)_{I_o=0}$, transmission parameter
C	capacitance
c	constant with a value between 1 and 2
D	$(I_i/U_o)_{U_o=0}$ transmission parameter
E	unspecified electrical quantity
E_c	controlled quantity
E_i	input quantity
E_ℓ	load quantity
E_s	source quantity
F	noise figure
f	frequency (in Hz)
f_ℓ	excess-noise corner frequency
f_T	current-gain-factor transit frequency
G	gyration constant
G	gain (unspecified), available power gain
g	transconductance
g_e	transconductance of a bipolar transistor
g_m	transconductance of a field-effect transistor
h_{FE}	dc current-gain factor of a bipolar transistor
I, i	current
K	constant in excess-noise spectra
k	Boltzman's constant ($1.38\cdot 10^{-23}$ JK^{-1})
k	feedback factor in the elementary feedback model

L	inductance
n	zero
n	number of devices (connected in series or in parallel)
n	transformer turns ratio
P	power
p	complex frequency
p	pole
Q, q	charge
q	charge of an electron ($1.6 \cdot 10^{-19}$ C)
R, r	resistance, differential resistance
S	power spectral density
T	temperature in K
T	transfer function (unspecified)
U, u	voltage
Y, y	(trans)admittance
Z, z	(trans)impedance
α	current-gain factor $(I_i/I_o)_{U_o=0}$
α_b	current-gain factor of a CB stage
α_e	current-gain factor of a CE stage
α_{eo}	low-frequency current-gain factor of a CE stage
β	feedback factor (parameter of the superposition model $(E_i/E_c)_{E_s=0}$)
γ	correlation factor
γ	transadmittance $(I_o/I_i)_{U_o=0}$
ε	differential error
ζ	transimpedance $(U_o/I_i)_{I_o=0}$
μ	voltage-gain factor $(U_o/U_i)_{I_o=0}$
ν	parameter of the superposition model $(E_\ell/E_c)_{E_s=0}$
ξ	parameter of the superposition model $(E_i/E_s)_{E_c=0}$
ρ	parameter of the superposition model $(E_\ell/E_s)_{E_c=0}$
τ	time constant
ω	angular velocity
ω_T	current-gain-factor transit frequency

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Design of High-Performance Negative Feedback Amplifiers

Amplifier design is very often regarded as making a selection from the large arsenal of known amplifier circuits and then adapting it for a specific purpose, possibly with the aid of computer-aided-design programs. Now and then designers are surprised by the introduction of a new amplifier circuit performing better in some respects than the others.

Each aspiring designer has to find his own way in this jungle. He has to choose from a rather chaotic and scattered collection of amplifier circuits rather than apply a systematic and straightforward design sequence that enables him to design his special-purpose amplifier circuit. A great deal of experience is essential.

This work is an attempt to make a useful contribution to the extensive literature on the subject of amplifier design. It can be justified on the grounds that the approach is believed to be unique in a number of respects. Many works that promise to cover the subject are instead concerned with analysis. Moreover, they frequently deal specifically or separately with particular design aspects, characterised by descriptions such as ‘wide-band’, ‘low-noise’, ‘low-distortion’, etc.

A treatment of the various design aspects and their interconnections, however, is necessary for fruitful amplifier design. At the basis of such a treatment lies the observation — usually easily overlooked — that amplifier design is concerned in the first place with obtaining an adequate quality of information transfer. Amplifiers are more than electronic circuits merely bringing the source power up to a higher level.

Quality requirements are imposed on the signal transfer relative to the type of information and to the manner of perception, registration, or processing. The quality of information transfer is determined by a large number of quality aspects such as linearity, accuracy, efficiency, signal-to-noise ratio, etc. Unfortunately, it cannot be expressed as a quantitative figure of merit.

A systematic, straightforward design approach is presented in this work. Preference was given to a qualitative rather than to a quantitative approach.

Finding the proper configurations for the basic amplifier and of the amplifier stages is considered of primary importance and is emphasised here. This book is therefore largely concerned with the design phase preceding the phase in which existing computer aids can be helpful.

Contents: Basic Amplifier Configurations for Optimum Transfer of Information from Signal Sources to Loads • A Classification and some Properties of Configurations with one Active Device • Design Considerations for Optimum Noise Performance of Negative-Feedback Amplifiers • Design Considerations Regarding Optimum Accuracy and Linearity of Negative Feedback Amplifiers • Design Considerations Regarding Optimum High-Frequency Performance of Negative-Feedback Amplifiers • The Design of Bias Circuitry • Outline of the Design Method • Appendix: Influence of the Induced Gate Noise in Wide-Band Amplifiers.

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